I. ZHEREBTSOV

FUNDAMENTALS OF RADIO

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РАДИОТЕХНИКА

ALEXANDER STEPANOVICH POPOV
(1859-1906)

Great Russian scientist, inventor of radio

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CHAPTER I

GENERAL INFORMATION ON RADIO COMMUNICATION

1. RADIO BROADCASTING AND COMMUNICATION

During the first years of its development, radio communication was called "wireless telegraph and telephone". However, this name was too long for. convenience and was later changed to "radio", which comes from the well-known Latin word "radius" — a straight line drawn from the centre of a circle to a point on its circumference. Wireless transmission was named radio transmission, or simply "radio" because radio stations emanate waves, like beams of light, in all radial directions or in certain directions.

The word "radio" now means the radiation of waves by transmitting stations, the propagation of these waves through space, and their reception by receiving stations. But this is not all. During its rapid growth art of radio has become closely associated with many other branches of science and engineering, and it is now difficult to limit the word "radio" to any simple definition.

Radio is sometimes defined as high-frequency engineering, but even this is not quite accurate, since it deals with both high-frequency and low-frequency currents. Defining radio as the technique of wireless transmission of electrical energy through space is also not entirely correct; it employs electrical energy to transmit sounds, images and telegraph signals, as well as special signals for radar, radio navigation and radio remote control.

Radio also includes the application of its techniques and equipment in other spheres, e.g., in medicine, biology, farming, metallurgy, machine-building, astronomy, geophysics, etc.

The major role in radio engineering equipment is that of various electronic devices. The branch of engineering dealing with the design and application of such devices is usually called "electronics". Hence, radio and electronics are developing in close association, supplementing each other and penetrating into numerous fields of science, engineering and culture. They are therefore frequently considered together and called radioelectronics.

The most developed application of radio is in communications and broadcasting. For several years after radio was invented it existed

only in the form of radio telegraphy. Then came radio telephony. Subsequent improvements have produced radio broadcasting, now serving millions of people.

A general diagram of a radio broadcasting system is shown in Fig. 1.

A microphone installed in a radio studio converts the sound waves of speech and music into alternating electric currents, known as *audio-frequency* or *low-frequency (LF) currents*, with a frequency

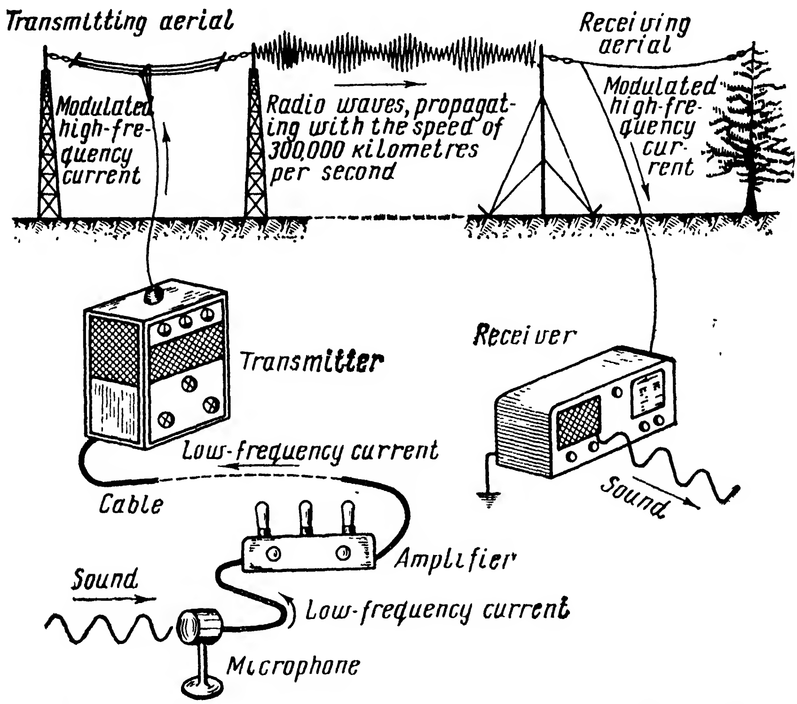


Fig. 1. A general scheme of broadcast transmission and reception

range extending from several dozens of cycles per second (cps) to approximately 10,000 cps. These currents are directed from the microphone to an amplifier, and the amplified low-frequency currents are then sent over a cable to a transmitting station, whose main apparatus is a radio transmitter. The transmitter is a device generating *high-frequency (HF)* or *radio currents* for the creation of radio waves. The frequency of these currents may be anywhere from 10,000 cps up to many millions and even milliards of cps.

The low-frequency currents arriving from the studio are superimposed upon the high-frequency current generated by the radio transmitter and change its amplitude in accordance with the changes of sounds in the programme. This process is called *modulation*.

The modulated high-frequency current is then fed by the radio transmitter to the aerial. The aerial emits radio waves, which travel in all directions with the speed of 300,000 kilometres per second (compare this with the speed of sound in the air, a mere 330 metres per second).

Under the influence of incoming radio waves a high-frequency current is set up in the aerial of a receiving station. This current, although very weak, is an exact replica of the current in the transmitting aerial and goes through the same changes, determined by the programme at the studio. The current picked up by the aerial is fed to a radio receiver, where it is first amplified and then converted into an audio-frequency signal. Such conversion is known as *detection* or *demodulation*. The audio-frequency current is then fed to a loudspeaker or earphones, where it is converted into sound waves.

One of the main properties of a radio receiver is its selectivity, i.e., its capacity to amplify only a narrow frequency band (to which it is tuned at the moment). Each transmitting station is assigned a wave (in effect, a narrow frequency band) different from those on which other stations operate. Owing to its selectivity, the receiver amplifies only currents of the wave which is assigned to the desired station. If selectivity were not incorporated in a radio receiver, the set would simultaneously amplify and reproduce the signals of all radio transmitting stations whose waves reach the receiving aerial.

The general diagram of a radio telegraph communication system is similar to that shown in Fig. 1 for radio telephony, the difference being that the telegraph station employs a key instead of a microphone. The key is so connected to the radio transmitter that it starts and stops the aerial radiation, which constitutes telegraph signals (dots and dashes), in space. At low-power stations (military portable stations, aircraft stations, etc.) the key and the microphone are located right at the station. In high-power installations the key and the microphone are usually placed at a considerable distance from the station and are connected to the transmitter by cables.

Transmission of signals by radio in one direction, i.e., from a transmitting station to a receiving station is known as one-way radio communication. Two-way communication calls for the installation of a transmitter, as well as a receiver, at each station.

Besides radio broadcasting, another widespread system is *sound diffusion* in which sound frequency waves from an amplifier located at a radio station are fed over wires to the loudspeakers of flats, clubs, etc. The radio station usually has a receiver which picks up the programmes of different stations and passes them on to subscribers via the amplifier. Some diffusion stations have their own sound studios for local transmissions. The diagram of a sound diffusion station is shown in Fig. 2.

Radio communication has the following important advantages over other types of communication:

(1) Possibility of contact at any time and over any distances.

(2) Practically instantaneous propagation of radio waves.

(3) Possibility of transmitting various types of information to any number of listeners.

(4) Transmission over any type of obstacle, e.g., oceans, seas, deserts, mountains, enemy-occupied territories, etc.

(5) Possibility of communication with moving objects (aircraft, ships, tanks, etc).

While possessing the above-listed advantages, radio communication has its own shortcomings, which are:

(1) Radio communication distance depends upon the time of day and year and also upon the wavelength.

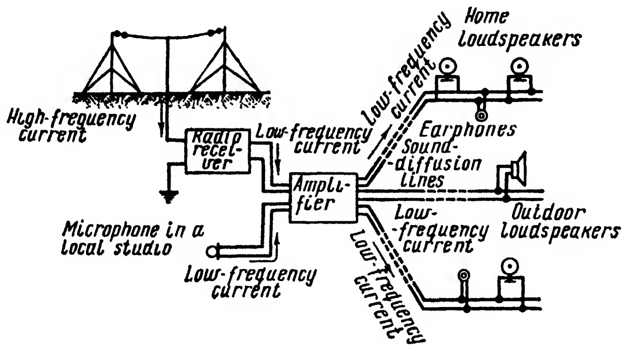


Fig. 2. A general scheme of a sound diffusion system

(2) Communication can be impaired by various types of electrical interference arising from atmospheric electrical discharges, various electrical installations, and other radio stations.

(3) Radio conversations can be listened to by anyone, and in order to secure secrecy, when required, special cyphers and other methods of privacy have to be used.

(4) Special radio direction finders can determine the location of a transmitting radio station.

2. WAVELENGTH

High-frequency alternating current is used to generate radio waves. Therefore the latter can be represented in terms of frequency, usually expressed in kilocycles per second (kc) or megacycles per second (mc.) Note that one kilocycle is equal to 1,000 cycles, while one megacycle is equal to 1,000,000 cycles.

Radio waves may also be expressed in terms of their length, i.e., in *wavelength*.

Let us first examine the meaning of wavelength on the surface of water. If we strike the water surface with a stick a number of

times we shall see that circular waves will propagate concentrically from the point of impact (Fig. 3). The faster we strike the surface of water with the stick, the greater will be the number of such circular waves and the shorter will become the distance between them. *The distance between the peaks of two succeeding waves* (or the distance between the two respective hollows) is known as the wavelength, represented by Greek letter λ in Fig. 3.

Wavelength is the distance travelled by the wave during one period, i.e., during one oscillation.

If the speed of radio wave propagation and the frequency are known, the wavelength can be determined. Assume, for instance, that the frequency of alternating current in the aerial of a radio station is 1,000,000 cycles per second (cps). In this case the oscillatory period is 0.000 001 of a second. Since the radio wave travels with the speed of 300,000 kilometres per second (300,000,000 metres per second), in 0.000 001 of a second it will travel one millionth the distance, i.e., 300 metres. This is the wavelength for the given set of conditions.

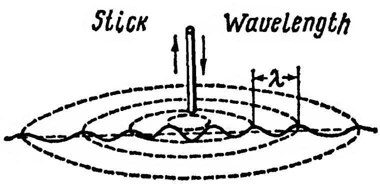


Fig. 3. Generation of water surface waves

Should the frequency of the alternating current in the aerial be halved and become 500,000 cps (500 kc), the oscillatory period will be doubled and equal to 0.000 002 of a second. During this length of time the radio wave will have travelled 600 metres instead of 300. Thus the higher the frequency, the shorter the wavelength, and vice versa.

The wavelength and the frequency are inversely proportional to each other.

Wavelength can be readily found by dividing the propagation speed of the wave (300,000 kilometres per second) by the frequency. To express the results in metres it is customary to represent the speed of propagation also in metres, or 300,000,000 m/sec. This gives the following two equations:

$$\lambda_{(\text{metres})} = \frac{300{,}000{,}000}{f_{(\text{cps})}}$$

$$f_{(\text{cps})} = \frac{300{,}000{,}000}{\lambda_{(\text{metres})}}.$$

If the frequency is expressed in kilocycles, the propagation speed should be given in kilometres per second (300,000), if the result is still to be expressed in metres, i.e.,

$$\lambda_{(\text{metres})} = \frac{300{,}000}{f_{(\text{kc})}}$$

$$f_{(\text{kc})} = \frac{300{,}000}{\lambda_{(\text{metres})}}.$$

Example 1. Find the wavelengths for frequencies of 600 and 300 kc.

Solution:

$$\lambda_1 = \frac{300,000}{600} = 500 \text{ metres}$$

$$\lambda_2 = \frac{300,000}{300} = 1,000 \text{ metres}$$

Example 2. A radio station works on 120 metres. Find its frequency.

Solution:

$$f = \frac{300,000}{120} = 2.500 \text{ kc}$$

3. RADIO WAVE RANGES

In accordance with agreements reached by International Conventions, radio waves are allocated to the following ranges, with the peculiarities noted below.

Long waves — waves with wavelengths between 30,000 and 3,000 metres (the respective frequencies are 10 to 100 kc). When radio was in its infancy, radio communication was based exclusively on long waves. Subsequent practice showed, however, that long-distance communication on such waves required radio transmitters of extremely high power. Moreover, the simultaneous operation of a large number of radio stations in the long-wave range was found to be impossible because of mutual interference, the nature of which is easily understood from the following. Clear broadcast transmission requires that each radio station is assigned an individual frequency spectrum (band) of approximately 9 kc. It can be readily shown that the whole of the long-wave range can accommodate only ten such bands, thus permitting only ten long-wave stations to operate simultaneously without mutual interference.

The only advantage of long waves is that they offer practically constant communication distance at any time of day and year. This cannot be attained by most waves of other ranges. At present only a small number of radio stations, sending time signals and weather reports operate in the long-wave range.

Medium waves — waves with wavelengths between 3,000 and 200 metres (their respective frequencies are from 100 to 1,500 kc). In this range, 2,000-200 m waves are specifically allocated to radio broadcasting and, accordingly, are of the greatest interest.

The 2,000-200 m sector of the medium-wave range is called the broadcast band. In practice this band is conventionally subdivided into "medium waves" (from 200 to 580 metres) and "long waves" (from 750 to 2,000 metres).

The simultaneous operation of as many as 150 broadcasting stations is possible in this range. Since this figure is considerably exceeded by the total number of radio broadcasting stations in Europe

alone, it becomes necessary to assign same wavelengths to several stations, which may bring about mutual interference. This can be alleviated to a considerable degree by assigning the same wavelengths to stations located at a large distance from each other.

Certain marine, aircraft and military radio telegraph stations are also assigned the 200-2,000 metre range, e.g., the 600-metre wave is assigned to marine communication and, in accordance with International Conventions, SOS signals are sent on this wave by ships in distress. Many ship and harbour stations use the waves of 580-750 metres for communication and, accordingly, these waves are not included in the broadcast bands.

The *short-wave* range extends from 200 to 10 metres (1.5-30 mc). Lower frequencies of this range (1.5-6 mc) are sometimes referred to as intermediate waves and are used for government telegraph and telephone radio communication.

Modern broadcast radio receivers are usually designed to tune over 13-50 and 200-2,000 metres. Short waves offer vast communication distances in comparison with other waves and attain this with relatively low transmitter power.

3,166 radio broadcast stations could operate simultaneously in the short-wave range of 200-10 metres if this range were solely allocated to radio broadcasting. A much greater number of radio telegraph stations could operate in the same range, as they require a much narrower individual band of frequencies.

The disadvantage of short waves is the very pronounced influence exerted upon them by the time of day and year. Nevertheless, the short-wave range is the most populated of all and is at present used by a vast number of various types of radio stations, including numerous broadcast and amateur stations all over the world.

Ultra-short waves (metre, decimetre, centimetre and millimetre waves) occupy the following ranges:

metre waves — from 10 to 1 metre (30-300 mc);
decimetre waves — from 100 to 10 cm (300-3,000 mc);
centimetre waves — from 10 to 1 cm (3,000-30,000 mc);
millimetre waves — from 10 to 1 mm (30,000-300,000 mc).

Waves shorter than 30 cm are sometimes referred to as microwaves.

Note: In certain countries the wavelength range is classified as follows:

V.H.F. — waves between 10 metres and 1 metre;
U.H.F. — waves between 1 metre and 10 centimetres;
S.H.F. — waves between 10 centimetres and 1 centimetre;
H.H.F. — waves between 10 millimetres and 1 millimetre.

Ultra-short waves, also called *ultra-high* and *super-high frequencies*, are used for communication between ground radio stations mainly

when the distance between the stations does not exceed 100-200 kilometres.

A vast number of radio stations can operate simultaneously in the ultra-short wave range without mutual interference. These waves are the only ones suitable for the transmission of television programmes. Ultra-short waves can be concentrated into a narrow beam, which can be trained in any desirable direction like the beam of a searchlight. This property of ultra-short waves is the basis of modern radar technique.

At present extensive research is being conducted in ultra-short wave engineering techniques, the stress being placed on the centimetre and millimetre waves and their application in communication, radiolocation, navigation aids and in other branches of science and engineering. Experiments are also being made with sub-millimetre waves with wavelengths of a fraction of a millimetre.

Chapter III gives more detailed information on the propagation of waves of various ranges.

4. QUESTIONS AND PROBLEMS

1. Which of the following are known as low-frequency and which as high-frequency currents: 300 kc; 8,000 cps; 30 mc; 150,000 cps; 6 kc; 1,250,000 cps; 425 cps?
2. What is the purpose served by high-frequency current at a transmitting radio station?
3. What will happen to the wavelength of a radio station if the frequency of current in its aerial is decreased three times?
4. A radio station operating on a wave of 250 metres shifts its wavelength to 1,500 metres. What will be the change of frequency of its aerial current?
5. Find the wavelengths corresponding to the following frequencies: 15 mc; 4,000 kc; 250 kc; 20,000 cps.
6. Find the frequencies corresponding to the following wavelengths: 6 metres; 50 metres; 375 metres; 1,200 metres; 25 centimetres.
7. To what ranges do the following waves belong: 115 metres; 243 metres; 49.5 metres; 3,506 metres; 481 metres; 15.5 centimetres; 31.4 metres; 6.7 metres; 84.1 metres?
8. To what ranges do the waves of the following frequencies belong: 5,300 kc; 12.6 mc; 249,500 cps; 1.5 mc; 187.5 kc; 38.7 mc; 8.25 mc; 2,730 kc?
9. Assuming that each station requires an individual frequency band of 9 kc, compute the number of free channels for broadcasting stations in the 200-580 metre medium-wave broadcast band and in the 750-2,000 metre long-wave broadcast band.

CHAPTER

OSCILLATORY CIRCUITS

5. FREE ELECTRICAL OSCILLATIONS

Circuits, in which high-frequency oscillations are set up, are the most important parts of all radio transmitters and receivers.

In order to have a clear understanding of the operation of such oscillatory circuits, let us first analyse the function of a pendulum comprised of a suspended weight performing mechanical oscillations (Fig. 4). If such a pendulum is pulled to the side and then released it will oscillate, swinging from position *1* to position *2* and back. Such oscillations are performed without the application of any external force, by virtue only of the initial energy stored in the pendulum when it was drawn aside. Such oscillations are known as free oscillations.

The movement of the pendulum from position *1* to position *2* and back represents *one complete oscillation*, or simply *one oscillation*. After the first oscillation comes the second, third, fourth, etc.

The maximum deflection of the pendulum from the zero position (*0*), i.e., distance *0-1* or *0-2*, is called the *amplitude of oscillation*.

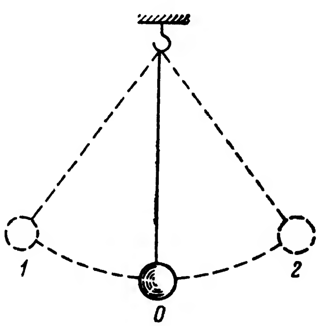


Fig. 4. Oscillations of a pendulum

The time taken by the pendulum to complete one oscillation is known as *a period* and is denoted by letter T. The number of oscillations per second is the frequency (f). The period is measured in seconds, frequency — in cycles (cps), kilocycles (kc) and megacycles (mc).

Free oscillations of a pendulum possess the following properties, easily checked by experiment:

(1) they are damped oscillations, i.e., their amplitude constantly decreases (goes through the damping process) as a result of energy losses occurring when the pendulum overcomes air resistance and the friction at the suspension point;

(2) free oscillations have a harmonic character, i.e., they are sinusoidal, if the damping effect is disregarded;

(3) the frequency of the free oscillations of the pendulum depends upon its length and is independent of the amplitude. As the damping process continues the amplitude of the oscillations decreases, but the period and frequency *remain constant;*

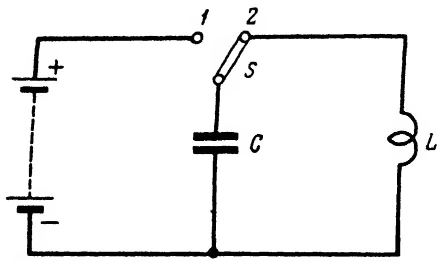


Fig. 5. A circuit for generating free oscillations

(4) the amplitude of the free oscillations is determined by the initial energy storage given to the pendulum when it is drawn aside. The longer the distance to which the pendulum is removed from its position of equilibrium, the greater will be the amplitude of oscillations.

As the pendulum oscillates the potential mechanical energy is converted into kinetic energy and back again. In position *1* or *2*, when the pendulum stops, it has the greatest potential energy, while its kinetic energy is equal to zero at that moment. As the pendulum swings from position *1* or *2* to *0*, the speed of its travel grows and the kinetic energy — the energy of motion — increases.

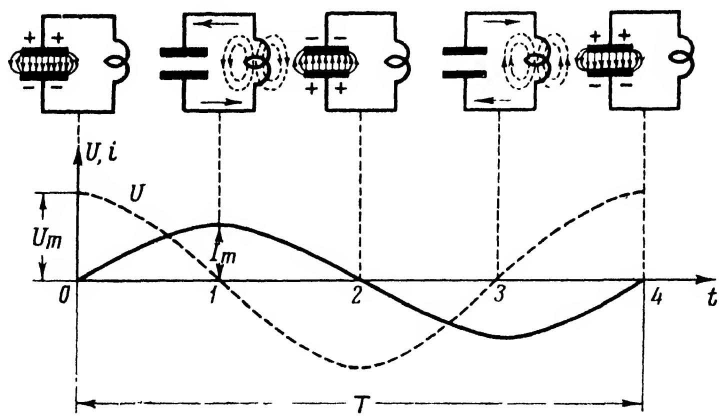


Fig. 6. The free oscillatory process in a closed capacitive-inductive circuit

When the pendulum passes through the *0* position, its speed and the kinetic energy have a maximum value, while its potential energy is equal to zero. When the zero point has been passed, the pendulum speed begins to decrease and the kinetic energy begins its conversion into potential energy. If there were no energy losses, the con-

version of energy from one state to the other would continue all the time and the oscillations would be continuous (undamped). But the energy losses are always present. Hence, if the pendulum is to oscillate continuously, it must be systematically given a push, i.e., periodically given additional energy, which will compensate for the losses. This is the principle of simple clockwork.

Let us now begin the study of electrical oscillations.

An oscillatory circuit represents a closed circuit consisting of inductance coil L and capacitor C. In the circuit diagram given in Fig. 5 such an oscillatory circuit is formed when change-over switch S is set to position *2*. Every oscillatory circuit naturally possesses some ohmic resistance, the effect of which we shall at present ignore.

The function of an oscillatory circuit is that of setting up, or generating, electrical oscillations.

If a charged capacitor is connected to a coil, the capacitor will discharge through the coil and the discharge will be of an oscillatory nature. To charge the capacitor, switch S is set to position *1* in the circuit diagram of Fig. 5. If the switch is then thrown over to contact *2*, the capacitor will begin to discharge through the coil.

The oscillatory process in this circuit can be conveniently traced by the graph of the changes of voltage u and current i in Fig. 6.

In the beginning the capacitor is charged to the greatest potential difference U_m, while the current i is at its zero value. As soon as the capacitor begins to discharge, a gradually increasing current will flow. The direction of electron movement of this current is shown by arrows in Fig. 6. The electromotive force set up by self-inductance of the coil (counter e.m.f.) will oppose any rapid changes of this current. As the current gradually increases, the voltage across the capacitor will decrease and at a certain moment (moment *1* in Fig. 6) the capacitor will be fully discharged. At such a moment the current will be maximum, while the voltage across the capacitor plates will be zero.

Such a situation — the presence of current in the absence of voltage — is easily explained as follows. The presence of inductance in the circuit prevents the current from sudden, instantaneous changes. The current is forced to decrease slowly and gradually by the counter e.m.f. set up in the coil owing to the change of current flowing through it. In this case, the coil acts as a generator and charges the capacitor. During this procedure the polarity of charge across the plates of the capacitor is altered. This is why the increasing voltage across the capacitor is represented in Fig. 6 by the down-sloping line directed towards the region of negative values.

When the charging of the capacitor is completed, the current will be zero and the voltage at maximum value, though the polarity of this voltage will be opposite to that shown for the initial moment (point *2* in Fig. 6). Hence after the completion of the pro-

cess described above, the polarity of the charged capacitor will have been reversed.

As the process continues the capacitor will again discharge through the coil, but in the opposite direction. Thereupon it will again be charged, and its polarity will be reversed after an exactly similar length of time. Thus the initial state of the oscillatory circuit will be restored at moment *4* of Fig. 6.

The electrons in the circuit will have completed one complete oscillation, the period of which is indicated by letter T in Fig. 6. This oscillation is then followed by a second, etc.

Free electrical oscillations are thus set up in the described circuit. These oscillations are really free because the oscillatory process goes on quite of its own accord and, owing to the initial charge of the capacitor, requires no application of any external electromotive force.

These oscillations are of a *harmonic* nature, i.e., they represent sinusoidal alternating current.

During the oscillatory process the electrons do not actually travel from one capacitor plate to the other. Although the current flow is extremely fast and approches the speed of light (300,000 kilometres per second), the electrons move in conductors very slowly—a mere few fractions of a centimetre per second. In one-half of a period the electrons can travel through only a small section of the conductor. They leave the negatively-charged capacitor plate and enter the closest portion of the connecting wire, while the other capacitor plate receives an equal quantity of electrons from the part of the conductor closest to the given plate. Thus in the conductors of an oscillatory circuit only a very slight shifting of electrons takes place.

A charged capacitor possesses a charge of potential electrical energy, which is concentrated in the electrical field between the capacitor plates. The movement of electrons is always accompanied by the creation of a magnetic field. Therefore the kinetic energy of the moving electrons is simply the energy of the magnetic field.

Electrical oscillation in a circuit is a repeated transfer of the potential energy of an electrical field into the kinetic energy of a magnetic field and back.

At first the entire energy is concentrated in the electrical field of a charged capacitor. When the capacitor begins to discharge, the potential energy begins to decrease, while the kinetic energy, the energy of the magnetic field of the coil, increases. When the current reaches its maximum and the whole of the energy of the circuit is concentrated in the magnetic field, the potential energy equals zero.

The process then continues in the reverse order; the magnetic energy decreases while the energy of the electrical field reappears. Half a period after the oscillation has begun the whole energy is again concentrated in the capacitor, which again leads to the transi-

tion of the energy of the electrical field into that of the magnetic field, etc.

The maximum value of current (or of the magnetic energy) corresponds to the zero value of voltage (or to the zero value of the electrical energy) and vice versa; i.e., the phase shift between the voltage and current is equal to one-quarter of a period, or 90°. During the first and the third quarters of a period the capacitor acts as a generator, while the coil is the energy consumer. The reverse takes place during the second and fourth quarters, when the coil works as a generator and feeds the magnetic field energy back to the capacitor, which now acts as the energy consumer.

An important feature of free oscillations in a circuit is *the equality of the inductive reactance of the coil and capacitive reactance of the capacitor in respect to the alternating current of the free oscillations.* This becomes clear from the following:

Capacitor and coil terminals are connected to each other; therefore the voltages across them are equal. The same current I flows through the capacitor and the coil, as the circuit being studied is a series circuit. This gives the following equation:

$$Ix_L = Ix_C,$$

where: x_L is the inductive reactance of the coil and x_C is the capacitive reactance of the capacitor. Dividing both parts of the equation by I, we have the following:

$$x_L = x_C.$$

This is the expression for the inductive or capacitive reactance possessed by the elements of the circuit at the frequency of natural oscillations and is referred to as the *characteristic impedance of a circuit*, denoted by Greek letter ϱ:

$$\varrho = x_L = x_C.$$

The value of ϱ in oscillatory circuits is usually equal to several hundred ohms.

6. AMPLITUDE AND FREQUENCY OF THE FREE OSCILLATIONS IN A CIRCUIT

The voltage and current amplitudes of free electrical oscillations in a circuit depend upon the initial energy storage. The greater the voltage of the initial charge of the circuit capacitor, the larger the amplitude of the oscillations.

Each oscillatory circuit is characterised by a very definite frequency of free oscillations, which is called *the natural frequency of an oscillatory circuit*, or simply *frequency* f_0 *of a tuned (oscillatory) circuit.*

Frequency f_0 depends upon the capacitance and inductance of the tuned circuit (C and L). The greater the capacitance and the inductance, the longer the period of oscillations and the lower their frequency.

If the capacitance is increased, the time taken for the capacitor to charge and discharge also increases, since at the former voltage the quantity of electricity (determining the charge) will be larger. Increasing the inductance of the coil will cause, in its turn, a slower increase and decrease of current during charging and discharging of the capacitor, since larger inductance offers greater opposition to current changes. Hence the circuit will oscillate more slowly, i.e., its frequency will be decreased. Decreasing L and C will cause the circuit to oscillate faster and the frequency will accordingly be increased.

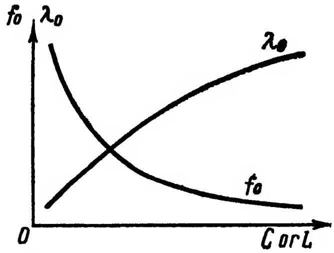


Fig. 7. Tuning curves of an oscillatory circuit

In order to decrease the frequency of a tuned circuit to one-half of its former value, the LC product must be increased four times. This can be obtained by a fourfold increase of either the inductance or the capacitance of the circuit. Another possibility is a twofold increase of capacitance and a similar simultaneous increase of inductance. Changing L or C, or their product, 9 times will change the frequency 3 times, etc.

It is possible to obtain the same frequency with different values of capacitance and inductance; but it is important that the product LC remains unchanged in the given case.

Wavelength is inversely proportional to frequency. Therefore any decrease of capacitance and inductance brings about a decrease of the wavelength (λ_0) of a tuned circuit, while any increase of capacitance and inductance increases the wavelength.

Fig. 7 illustrates the curves representing the natural frequency f_0 of a tuned circuit and corresponding wavelength λ_0, and the dependence of these parameters upon the value of capacitance or inductance contained in the circuit. These curves are known as the *tuning curves.*

Thomson's formula gives the dependence of the tuned circuit frequency upon the circuit inductance and capacitance:

$$f_0 = \frac{1}{2\pi\sqrt{LC}} = \frac{1}{6.28\sqrt{LC}}.$$

Here, frequency f_0 is expressed in cycles per second, L — in henries and C — in farads.

English scientist Thomson was the first to give the formula for the period of free oscillations in a tuned circuit:

$$T = 2\pi\sqrt{LC}.$$

Modern radio engineering, however, prefers to express oscillations in terms of frequency, because the period represents a very small fraction of a second, and this is inconvenient in calculations.

Thomson's formula is easily derived from the equality of inductive and capacitive reactances. As we already know, $x_L = x_C$ in all cases when free oscillations take place in a tuned circuit, i.e.,

$$2\pi f_0 L = \frac{1}{2\pi f_0 C}$$

or:

$$f_0^2 = \frac{1}{4\pi^2 LC}.$$

Hence:

$$f_0 = \frac{1}{2\pi\sqrt{LC}}.$$

In any oscillatory circuit the frequency of free oscillations depends upon two parameters. In electrical oscillatory circuits these parameters — inductance and capacitance — are easily changed. In case of the common pendulum (see Fig. 4) one parameter — the length of the pendulum — can be also changed, and it can be shown that the length has to be changed fourfold to cause a twofold change in frequency, ninefold to cause a threefold frequency change, etc. The other parameter of the pendulum is the acceleration caused by gravitation. This value is expressed by $g = 9.81$ m/sec^2 and cannot be varied.

The best mechanical analogy of an oscillatory circuit is the spring pendulum shown in Fig. 8. The frequency of natural oscillations of such a pendulum depends upon the mass of the suspended weight and the flexibility of the spring. Flexibility is inversely proportional to resilience and denotes the degree to which the spring is prone to stretching and compression when acted upon by the applied force. The value of flexibility depends upon the thickness and material of the spring, as well as upon the diameter of spring turns and their number. If the number of turns of the spring is increased 4 times, the flexibility will be also increased fourfold, while the frequency of oscillations will be reduced twofold. The same frequency change will be obtained by changing the weight fourfold. Hence the dependence of the frequency of free oscillations upon the two parameters is easily shown with the help of the given type of pendulum.

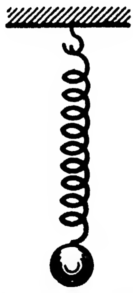


Fig. 8. Spring pendulum

7. DAMPED AND CONTINUOUS OSCILLATIONS

Thus far we have been analysing a perfect tuned circuit, consisting only of capacitance and inductance, reactive in character and causing no energy losses. When there is no resistance in the circuit, the amplitude of oscillations remains constant and they can continue indefinitely. Such oscillations are called *continuous* (Fig. 9*a*).

In reality, every oscillatory circuit possesses a certain amount of resistance (otherwise called the ohmic resistance). This resistance

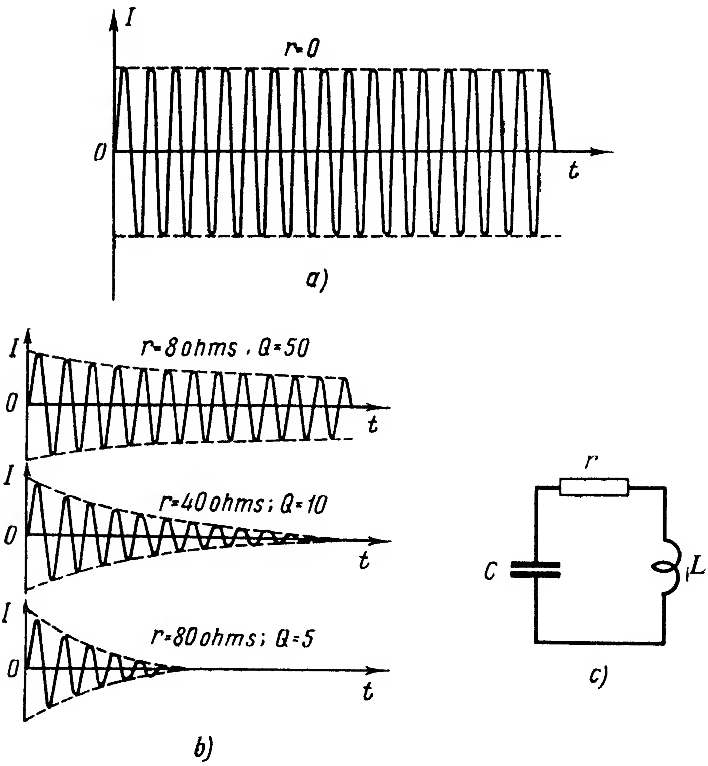


Fig. 9. Continuous (*a*) and damped (*b*) oscillations. Equivalent circuit of a practical oscillatory tuned circuit (*c*)

is mainly distributed in the coil, partly in the connecting wires and also in the capacitor. Fig. 9*c* shows the so-called equivalent circuit of a practical oscillatory tuned circuit, where the total ohmic resistance r is conventionally shown connected in series, and the coil and capacitor are supposed to contain no resistance. The ohmic resistance is sometimes referred to as *the loss resistance*.

The following types of high-frequency current losses occur in a tuned circuit:

(1) Losses caused by the heating of the conductor which, as a result of the skin-effect, has higher resistance on high frequency than it does to direct current. The skin-effect is caused by the flowing of the high-frequency current over a thin surface layer of the conductor rather than through its entire cross-section. As a result, the effective cross-section of the conductor is decreased, which increases its resistance. The higher the frequency, the thinner the surface layer carrying the high-frequency current and the higher the conductor resistance.

(2) Losses caused by the heating of solid dielectrics, in which the alternating electrical field produces oscillation of molecules accompanied by friction, which causes losses due to dielectric hysteresis.

(3) Losses caused by leakage currents, appearing because solid dielectrics are not ideal insulators. (At high voltages leakage of electrical charges into the air is also observed.)

(4) Losses caused by the heating of the ferromagnetic cores used to increase the inductance of coils. This heating is caused by magnetic hysteresis and eddy currents in the cores.

(5) Eddy current losses in all metal objects in the proximity to the oscillatory circuit and within the influence range of the alternating magnetic field.

(6) Losses caused by the radiation of electromagnetic waves by the oscillatory circuit.

(7) Losses caused by the pickup of high-frequency energy by other circuits coupled with the given oscillatory circuit.

Almost all these types of losses increase with an increase of frequency.

All these losses may be regarded as equivalent to the losses in a certain ohmic resistance, denoted by letter r. Thus, the ohmic resistance of a tuned circuit expresses the total energy losses occurring in the given tuned circuit.

It is the ohmic resistance which causes the damping of oscillations. Owing to the presence of such resistance in a tuned circuit, the amplitude of oscillations is gradually reduced and within a short period of time becomes so small that the oscillations are considered non-existent.

Free oscillations in a practical tuned circuit are always of a damped character.

The larger the ohmic resistance of a tuned circuit, the greater is its damping effect upon the oscillations. Fig. 9*b* gives the curves of damped oscillations of the same tuned circuit for different values of ohmic resistance. The frequency of oscillations remains constant from beginning to end, in spite of the gradually decreasing amplitude. If the value of the ohmic resistance in a tuned circuit is very large,

the damping effect becomes so pronounced that the circuit refuses to oscillate at all.

The ohmic resistance affects the frequency of oscillations to some extent. The greater the ohmic resistance, the lower the frequency. However, this influence is so insignificant that it is disregarded in practical applications.

From a mathematical point of view, the value of damping of the oscillations is generally considered as the ratio of ohmic resistance r and characteristic impedance ϱ. This ratio is known as oscillatory circuit damping and is denoted by Greek letter δ:

$$\delta = \frac{r}{\varrho} = \frac{r}{2\pi f_0 L}.$$

The greater r in comparison with ϱ, the greater the damping. In well-designed tuned circuits δ is less than 0.01. Average tuned circuits have δ values between 0.05 and 0.01. If the δ value exceeds 0.05, the quality of the tuned circuit is considered to be poor.

The quality of tuned circuits is also evaluated on the basis of *quality factor* inversely proportional to damping. The quality factor is usually represented by letter Q and is equal to the ratio of the characteristic impedance to the ohmic resistance of a tuned circuit.

$$Q = \frac{1}{\delta} = \frac{\varrho}{r} = \frac{2\pi f_0 L}{r}.$$

The smaller the damping of a tuned circuit, the higher is its Q. The value of Q in an average tuned circuit is between 20 and 100. If Q exceeds 100, the tuned circuit is considered to be of high quality. Poor tuned circuits have Q values of less than 20.

Modern radio communication requires continuous oscillations, sustained over any length of time. Such oscillations can be produced by periodical application of energy to a tuned circuit to compensate the circuit losses.

In practice this is obtained by connecting the tuned circuit to a e.m.f. source, e.g., a battery, which will replenish the charge of the capacitor. Such connection should be made with a frequency equal to that of the tuned circuit and only during those quarters of a period when the capacitor is charged. Obviously such a procedure cannot be carried out manually, nor can it be performed automatically with the help of an electromagnetic relay, for the latter always has considerable inertia. At frequencies of hundreds of thousands and millions of cycles per second, only electronic valves or semiconductors can serve as the required automatic relays. The design and operation of electronic valves and their functions in maintaining continuous oscillations in a tuned circuit are studied in Chapter IV.

8. FORCED OSCILLATIONS AND RESONANCE

Unlike free oscillations, forced oscillations do not take place of their own accord but are set up by some periodic external force. For example, electrical oscillations in a receiving aerial are not free, for they are not caused by an initial storage of energy but are created by incoming radio waves.

Let us first analyse the forced oscillations of a pendulum possessing a definite natural frequency. Let us swing this pendulum manually at a different frequency. The character of this swing (oscillation) depends upon the movement of our hand and may be, in particular, sinusoidal. External energy is periodically applied to the pendulum; hence its oscillations will be continuous and may possess any frequency, determined only by the frequency of the applied external force.

A similar phenomenon may also take place in an electrical oscillatory circuit if it is connected to a generator of alternating current. At any frequency of the generator, alternating current will flow through the tuned circuit, i.e., forced electrical oscillations will take place in the tuned circuit at the generator frequency.

Forced oscillations possess properties quite different from those of free oscillations. They may be summarised as follows:

(1) forced oscillations are continuous and are not subjected to damping (more precisely, they exist as long as the external e.m.f. is applied to the tuned circuit);

(2) they may have different forms, depending upon the character of the e.m.f.;

(3) their frequency does not depend upon L and C of the tuned circuit but is determined by the frequency of the applied e.m.f.;

(4) their amplitude depends not only upon the value of the applied e.m.f. but also upon the relation between the frequency of this e.m.f. and the natural frequency of the tuned circuit.

The last listed peculiarity is of special interest and must be analysed in detail.

Free oscillations of a definite frequency take place in any tuned circuit which has been given a storage of energy. When the damping is not very pronounced, even this small initial storage of energy will sustain the oscillations for considerable time. Now, as we already know, external periodic e.m.f. has to be applied to the circuit in order to sustain indefinitely oscillations of the forced nature. The greater the difference between the natural frequency of the circuit and the frequency of the applied e.m.f., the higher must be the value of this e.m.f. to sustain oscillations. The smaller the difference of the frequencies, the greater will be the amplitude of the forced oscillations and the smaller will be the energy required to sustain them. If the frequency of the applied e.m.f. is equal to the natural frequency of the tuned circuit, the amplitude of the oscillations will be at its

maximum and the application of only an insignificant amount of energy will sustain the oscillatory process. Such a state is referred to as *resonance.*

The phenomenon of resonance obtains when, with the coincidence of the frequency of externally applied e.m.f. and the natural frequency of the tuned circuit, the amplitude of forced oscillations reaches the maximum value.

Equality of the external generator and tuned circuit frequencies is the condition required for resonance.

Of course, such equality of frequencies is a purely mathematical condition of resonance, which should always be considered a certain process characterised by definite properties.

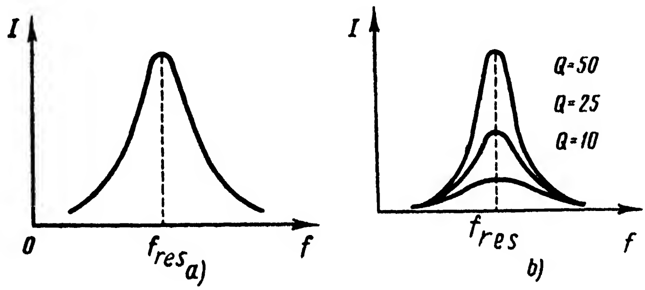


Fig. 10. Resonance curves of a tuned circuit

The frequency at which resonance occurs is called the *resonant frequency* or f_{res}: it is equal to the natural frequency f_0 of the tuned circuit.

The condition of resonance may be obtained either by changing the generator frequency (while keeping the natural frequency of the tuned circuit constant) or by changing the natural frequency of the tuned circuit, by varying C or L, while the generator frequency remains constant. In both cases resonance is clearly illustrated by the curves in Fig. 10*a*, showing the depedence of the amplitude of oscillations in a tuned circuit upon the frequency. Variable frequency (either of the generator or of the tuned circuit) is marked off along the X-axis, which also shows the value of f_{res}. The current through the tuned circuit is marked off along the Y-axis. As may be seen from the resonance curve, when the condition of resonance exists the current in the tuned circuit (i.e., the amplitude of the forced oscillations in the circuit) is at the maximum value. It may also be seen that the circuit current decreases when the frequency changes in either direction.

The damping effect of the tuned circuit strongly influences the resonance. If the resonance curves are plotted to the same scale for several tuned circuits possessing various degrees of damping,

it will be observed that in the circuits with smaller damping the resonance curves will be sharper and more pronounced. This is clearly shown in *Fig. 10b*. A well-defined resonance curve is an indication that the tuned circuit has practically no response to oscillations differing in frequency from the natural frequency of the circuit, but at resonance with the right frequency such a circuit will readily respond, and strong oscillations will be set up in it. This is called *sharp resonance*. In a heavily-damped circuit, the amplitude of oscillations at resonance is low and the circuit will respond to oscillations whose frequencies considerably differ from the resonance frequency. This is called a *broad resonance*.

The smaller the damping in a circuit, the sharper is the resonance and the higher is the circuit sensitivity to oscillations of the resonance frequency.

The condition of resonance is produced by obtaining powerful high-energy oscillations by very little external excitation. This excitation is needed only for the compensation of energy losses occurring in the circuit in its oscillatory state.

9. SERIES RESONANCE

Two different kinds of resonance occur in tuned circuits, *series resonance* and *parallel resonance*.

Series resonance is observed when the generator supplying the alternating e.m.f. is connected to a load consisting of series-connected L and C of a tuned circuit, which places the generator itself in series with the coil and capacitor as shown in Fig. 11*a*.

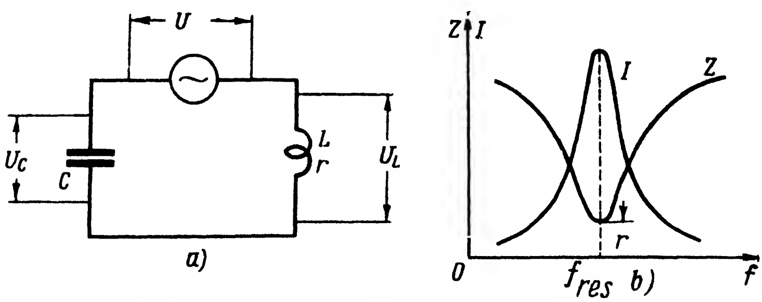


Fig 11. Series resonance curves and circuit

A circuit of this type has a certain amount of ohmic resistance r and total reactance x, the latter being equal to the difference of inductive and capacitive reactances

$$x = x_L - x_C.$$

We consider the difference of inductive and capacitive reactances because each affects the current in the opposite way. The first type

of reactance makes the current lag behind the voltage, while the other one makes the current lead the voltage.

On natural resonance x_L and x_C are equal to each other. If the *generator frequency is equal to the natural frequency of the tuned circuit,* then x_L is also equal to x_C for the current produced by the generator. In this case the total reactance x becomes zero and the impedance of the circuit as presented to the generator will be equal only to the ohmic resistance, which is comparatively low in tuned circuits. This affect accounts for the considerable current increase in the circuit at resonance and eliminates the phase shift between current and voltage.

In a series-resonant circuit the impedance of the circuit decreases to its minimum value at resonance, equal to the ohmic resistance of the circuit, while the current flowing through the circuit reaches its maximum value.

The conditions of obtaining series resonance require an equality of the generator frequency and the natural frequency of the tuned circuit, or an equality of inductive and capacitive reactances as presented to the generator.

When the generator frequency is higher than the frequency of the tuned circuit, the inductive reactance predominates over the capacitive reactance, and the tuned circuit presents some inductive reactance to the generator.

If, however, the generator frequency is lower than the frequency of the tuned circuit, the capacitive reactance will be greater than the inductive reactance, and the tuned circuit will present some capacitive reactance to the generator.

Since in both cases the frequency deviates from its resonance value, the impedance of the tuned circuit will become greater than it was during resonance.

Fig. 11*b* illustrates impedance change and current change curves plotted for changing generator frequency, where Z denotes impedance, I — current and f — generator frequency.

The following simple relations are used in calculations of impedance of tuned circuits:

$$Z = r; \qquad I = \frac{U}{r}.$$

From this it follows that the generator voltage U is equal to the voltage drop across the ohmic resistance r.

Heavy current circulating in the tuned circuit during resonance builds up voltages across the inductive and capacitive reactances; these voltages will be considerably greater than the generator voltage, and equal to

$$U_L = Ix_L; \qquad U_C = Ix_C.$$

Since $x_L = x_C = \varrho$, these voltages are equal to each other but opposite in phase and, consequently, compensate each other. This is because the current has the same value in all parts of the circuit; *the voltage across the coil leads the current by 90°, while the voltage across the capacitor lags behind the voltage by 90°*. Therefore under such circumstances the phase shift between these voltages is 180°.

The resonance curve of the current given in Fig. 11*b* also illustrates the change of voltage U_L or U_C (only on a different scale) for small frequency changes. This is easily understood if we consider that when the frequency changes near its resonance value, the current value changes sharply while the values of x_L and x_C change relatively little. For instance, if f_{res} is 1,000 kc and the frequency shifts over 20 kc, i.e., changes by 2%, each of the two reactances, x_L and x_C, will also change by 2%. As a result, voltages $U_L = Ix_L$ and $U_C = Ix_C$ will change almost in exact proportion to the current.

It can be readily shown that when the circuit is in the state of series resonance, the voltage across the coil or across the capacitor is equal to the generator voltage multiplied by Q. The generator voltage U is equal to Ir. The voltage across L or across C may be represented as $U_C = U_L = I\varrho$.

Hence:

$$\frac{U_L}{U} = \frac{I\varrho}{Ir} = \frac{\varrho}{r} = Q.$$

The greater the value of Q, the higher is the voltage increase at resonance.

Voltage increase across the coil and capacitor is a typical effect of series resonance. This is why series resonance is sometimes referred to as "voltage resonance", to emphasise the increase of voltages during the resonant condition of the circuit.

High voltages appearing across the coil and capacitor are accounted for by the gradual storage of energy in the tuned circuit during the oscillatory process. The generator e.m.f. excites the tuned circuit, setting it into oscillation, and the amplitude of these oscillations increases until the energy supplied by the generator becomes equal to the energy losses in the ohmic resistance of the circuit. This stage attained, powerful oscillations appear in the tuned circuit, these oscillations being associated with heavy current and high voltage, while the generator delivers a comparatively low power for the sole purpose of compensating the energy losses.

This condition may be likened to the following case with a heavy pendulum, when the latter is made to swing back and forth by light hand motions applied in time with the natural oscillation frequency of the pendulum, until the swing (oscillation) amplitude of the device gradually increases and finally begins considerably to exceed the amplitude of the swinging hand, which, in the given case, may be likened to the electrical generator used to excite the tuned circuit.

Series resonance is applied in radio engineering when it is required to obtain maximum current and voltage values in a tuned circuit.

For instance, the aerial tuned circuit of a radio transmitter is a series-resonant circuit, which permits the transmitter to "pump" the greatest possible amount of current into the aerial, securing the longest communication range possible with the given type of transmitter.

The input circuit of a radio receiver is also frequently represented by a series-resonant tuned circuit, which makes possible the obtaining of voltage amplification in the aerial and the selection of radio signals transmitted by a radio station. Owing to the series resonance effect, the tuned aerial circuit of the receiver amplifies the signal of only that transmitting station to whose frequency it is tuned. The signals of other stations, working on different frequencies are amplified only very slightly in the input circuit of such a receiver.

In dealing with the series resonance it must be kept in mind that the internal resistance of the generator is a part of the ohmic resistance of the tuned circuit. If the generator has a high internal resistance, the quality of the tuned circuit may be impaired and its resonance properties insufficiently pronounced. Therefore the generator used for feeding series-resonant circuits with alternating current must be of the low-resistance type.

Let us analyse the phenomenon of series resonance by means of a numerical example.

Given a series-resonant tuned circuit with the following parameters: $L = 4$ millihenries, $C = 160$ picofarads, $r = 50$ ohms, 25-volt generator is used to supply this circuit with alternating current. Determining the natural frequency of the tuned circuit, we have:

$$f_0 = \frac{1}{2\pi\sqrt{4\times10^{-3}\times160\times10^{-12}}} \approx \frac{10^7}{6.25\times8} = 2\times10^5 \text{ cps} = 200 \text{ kc}.$$

To simplify our calculations, let us assume that 2π is equal to 6.25 (instead of 6.28). Next, we shall find the values of x_L and x_C:

$$x_L = 6.25 f_0 L = 6.25\times2\times10^5\times4\times10^{-3} = 5{,}000 \text{ ohms},$$

$$x_C = \frac{1}{6.25 f_0 C} = \frac{10^{12}}{6.25\times2\times10^5\times160} = 5{,}000 \text{ ohms}.$$

Thus we have shown that $x_L = x_C = \varrho$.

The quality factor for the given circuit is:

$$Q = \frac{\varrho}{r} = \frac{5{,}000}{50} = 100.$$

If the generator frequency is 200 kc, the phenomenon of series resonance will occur. Then $Z = r = 50$ ohms and the tuned circuit current will be given by the following expression:

$$I = \frac{U}{r} = \frac{25}{50} = 0.5 \text{ ampere}.$$

Voltages across L and C are next determined:

$$U_L = U_C = IQ = 0.5 \times 5{,}000 = 2{,}500 \text{ volts.}$$

These voltages are 100 times greater than the generator voltage (which is logical, because $Q = 100$).

Now we can determine the oscillatory power, i.e., the reactive power in the coil and capacitor:

$$P_L = P_C = I^2 Q = 0.5^2 \times 5{,}000 = 1{,}250 \text{ volt-amperes.*}$$

Continuing our calculations, we next determine the power expended by the generator for the purpose of sustaining the oscillations. This is the active power lost in resistance r and it is equal to:

$$P = I^2 r = 0.5^2 \times 50 = 12.5 \text{ watts.}$$

This active power is 100 times less than the reactive oscillatory power.

Should the generator frequency change, e.g., increase by 10% and become equal to 220 kc, the condition of resonance will be upset; x_L will increase by 10% and will be 5,500 ohms, while x_C will decrease by 10% and will be 4,500 ohms. The total reactance under these circumstances will be equal to $x = x_L - x_C = 5{,}500 - 4{,}500 = 1{,}000$ ohms and will be of an inductive character. In this case the impedance of the tuned circuit may be said to be equal to its reactance, because $Z = \sqrt{r^2 + x^2} = \sqrt{50^2 + 1{,}000^2} \approx 1{,}000$ ohms.**

In comparison with the impedance of 50 ohms, possessed by the tuned circuit at resonance, this impedance has increased 20 times, dropping the current an equal amount. The current value will now be:

$$I = \frac{25}{1{,}000} = 0.025 \text{ ampere.}$$

A similar result would have been obtained if the generator frequency had dropped below the resonance value instead of rising above it.

10. PARALLEL RESONANCE

Parallel resonance is obtained when inductance and capacitance generators are in parallel, i.e., when the generator is connected outside the circuit (Fig. 12*a*).

However, in analysing the oscillatory circuit made up by the inductance and capacitance apart from the generator, it should still be considered as consisting of series-connected L and C. It would be wrong to say that in this type of arrangement the oscillatory circuit is connected in parallel with the generator. The circuit, taken as a whole, serves as the generator load resistance and, consequently, it is connected with the generator in series—which is normal in any closed circuit.

In a parallel-resonant circuit the conditions for obtaining resonance are the same as in a series-resonant circuit, i.e.: $f = f_0$ or

* Remember that the reactive power is measured in volt-amperes; a watt is a measurement unit of active power.

** When an ohmic resistance and a reactance are connected in series and one of them is 10 or more times larger than the other, the impedance of the circuit can be assumed roughly equal to the greater of the two values.

$x_L = x_C$. It should be noticed, however, that parallel resonance differs in many respects from series resonance. In the type of circuit now being studied, voltage across the coil-capacitor combination and across the capacitor or the coil is the same as the generator voltage. At resonance the impedance across the tuned circuit (across the combination) becomes maximum and the generator current—minimum (which is quite the opposite to what is observed during series resonance). The equivalent resistance of the tuned circuit presented to the generator can be found from any one of the following formulas on parallel resonance (R_e):

$$R_e = \frac{\varrho^2}{r} = Q\varrho = \frac{L}{rC},$$

where: L is expressed in henries, C — in farads, while R_e, ϱ and r — in ohms.

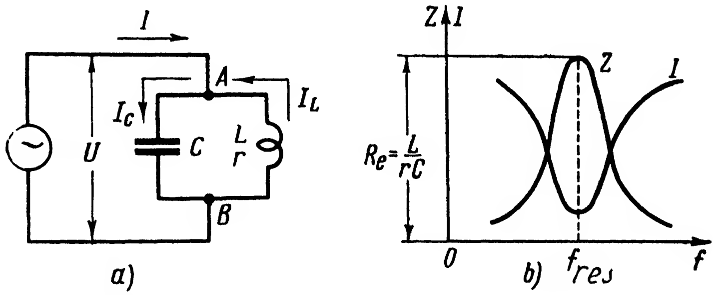


Fig. 12. Parallel resonance curves and circuit

R_e, otherwise called *resonant resistance*, is a pure ohmic resistance, which accounts for the fact that there is no phase shift between current and voltage of the generator under conditions of parallel resonance.

Fig. 12*b*, plotted for a case of parallel resonance, shows how the impedance Z of the tuned circuit and the generator current I change with the change of the frequency f of the generator.

At resonance strong oscillations take place in the closed circuit formed by L and C; therefore, the circulating current through the coil and capacitor is much larger than the current supplied by the generator. The currents flowing through the coil (I_L) and capacitor (I_C) may be considered as currents through the branches of an electrical circuit. They may also be regarded as the current of continuous oscillations in a tuned circuit, the oscillations being sustained by the generator.

In respect to voltage U, current through the coil lags by 90°, and current through the capacitor leads by 90°. This means that the two currents are shifted in phase by 180° in respect to each other. Owing to the presence of ohmic resistance, concentrated mainly

in the coil, currents I_C and I_L actually have a lesser phase shift than 180°, and the current through the coil is slightly less than that through the capacitor. Hence, in compliance with Khirhoff's first law, the following equation may be written for the branching-off point:

$$I + I_L = I_C,$$

or

$$I = I_C - I_L.$$

The smaller the value of ohmic resistance in a tuned circuit, the smaller is the difference $I_C - I_L$, the smaller the generator current, and the greater the tuned circuit resistance. This is quite logical. Current supplied by the generator replenishes the energy in the tuned circuit, making up for its losses in ohmic resistance. If the ohmic resistance is decreased, the losses which occur in it are also decreased and less power has to be taken from the generator to sustain continuous oscillations in the tuned circuit. If the tuned circuit were an ideal one—a circuit introducing no losses—then the oscillations, once commenced, would continue indefinitely without any damping, and no power at all would be required from the generator.

The generator current in this case would be equal to zero and the impedance of the tuned circuit would be infinity.

The active power delivered by the generator may be calculated as follows:

$$P = IU = I^2R_e = \frac{U^2}{R_{oc}}.$$

Alternatively it may be computed as power lost in the ohmic resistance of the tuned circuit:

$$P = I_{oc}^2 r,$$

where I_{oc} is the current in the oscillatory circuit, being equal to I_L or I_C.

*Parallel resonance, like series resonance, is characterised by setting up powerful oscillations in the tuned circuit, calling for only insignificant consumption of energy from the generator.**

In a parallel-resonant circuit the internal resistance (R_i) of the generator exerts considerable influence. If the generator resistance is low, the voltage at the generator terminals, and hence the voltage across the tuned circuit, differs only slightly from the e.m.f. of the generator. Under such circumstances the voltage amplitude across the tuned circuit remains practically constant as the current changes with the frequency. This is proved by the following. We know that

* In this process the power of oscillations in the tuned circuit is, of course, reactive power.

$U = E - IR_i$. In this case, however, the value of R_i is so small *that the voltage drop IR_i in the generator is insignificant* and U may be considered practically equal to E.

In the circuit being studied the impedance of the whole circuit is practically expressed by the tuned circuit impedance. The latter sharply increases at resonance, causing a sharp decrease in the generator current. This condition is shown by the current change curve in Fig. 12*b*.

The unchanging voltage amplitude across the tuned circuit can be also explained on the basis of formula $U = IZ$. In resonant condition Z has a high value, but I is low, and if the condition of resonance is not present, the value of Z decreases. However, I increases and the product IZ remains approximately unchanged.

As may be seen, when the internal resistance of the generator is low a parallel-resonant circuit possesses hardly any resonant properties in respect to voltage, i.e., the voltage across the tuned circuit hardly increases at resonance. Therefore, currents I_L and I_C will not noticeably increase on resonance, and when the generator internal resistance is low, a parallel-resonant circuit will not have any pronounced properties in respect to currents flowing through the coil and capacitor.

In practical radio equipment a parallel-resonant circuit is generally supplied with power from high-resistance generators, such as an electron valve or a semiconductor circuit. If the internal resistance of a generator is considerably in excess of the impedance value Z of a tuned circuit, the parallel-resonant circuit acquires sharply pronounced resonant properties.

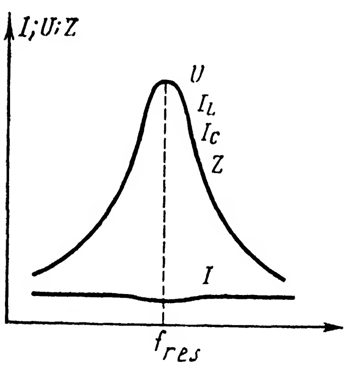


Fig. 13. Resonance curves of a parallel tuned circuit when the internal resistance of the generator is high

In this case Z of the complete circuit is approximately equal to R_i and remains unchanged even though the frequency may be changing. Current I, feeding the tuned circuit, will likewise remain practically unchanged in amplitude and may be given by the following equation:

$$I = \frac{E}{R_i + Z} \approx \frac{E}{R_i}.$$

But then the voltage across the tuned circuit ($U = IZ$) will follow impedance changes of the tuned circuit as the frequency changes, i.e., U will increase sharply at resonance. Currents I_L and I_C will also increase correspondingly. Thus when the internal resistance of the generator is high, the curve showing the impedance Z change in Fig. 12*b* will also indicate, on a different scale and approximately,

the change of voltage U across the tuned circuit and changes of currents I_L and I_C. Fig. 13 illustrates this type of curve together with the generator current change curve, the latter showing that the current in this case hardly varies.

The main application of parallel resonance in radio is that of creating high load impedance for current of a definite frequency in valve oscillators and high-frequency amplifiers.

Let us now consider a numerical example of parallel resonance. For the purpose of comparison with series resonance we shall take the tuned circuit used in the previous example, connecting this circuit in accordance with the schematic diagram in Fig. 12 and using the same generator with an output voltage of 25 volts.

The impedance of the tuned circuit (R_e) is equal to:

$$R_e = Q\varrho = 100 \times 5{,}000 = 500{,}000 \text{ ohms.}$$

Generator current is:

$$I = \frac{U}{R_e} = \frac{25}{500{,}000} = 0.00005 \text{ a} = 0.05 \text{ ma.}$$

Tuned circuit current $I_{oc} = I_L = I_C$ will be equal to:

$$I_{oc} = \frac{U}{x_L} = \frac{25}{5{,}000} = 0.005 \text{ a} = 5 \text{ ma.}$$

The resonant impedance of the tuned circuit (R_e) is 100 times as great as the reactance of the coil or capacitor at the same frequency, hence, the generator current will be 100 times less than the current in the tuned circuit. The same relation will be established between power levels. The reactive power of oscillations in the tuned circuit is equal to:

$$P_L = UI_L = 25 \times 0.005 = 0.125 \text{ volt-ampere,}$$

while the active power drawn from the generator is:

$$P = UI = 25 \times 0.00005 = 0.00125 \text{ watt} = 1.25 \text{ milliwatts.}$$

Alternatively, this power may be computed as follows:

$$P = I_{oc}^2 r = 0.005^2 \times 50 = 0.00125 \text{ watt} = 1.25 \text{ milliwatts.}$$

Let us also take an example comparing the operation of a parallel-resonant circuit for different values of internal resistance of the generator.

Assume a parallel-resonant circuit in which R_e at resonance is equal to 10,000 ohms and $x_L = x_C = \varrho = 200$ ohms. Let $Z = 1{,}000$ ohms on slight detuning. When this tuned circuit is supplied with power from a generator with $E=2$ volts and $R_i = 0$, then the voltage across the circuit is always equal to $U=E=2$ v, whether the circuit is tuned to resonance or not. Since x_L and x_C vary very little on slight detuning, it may be assumed that the currents through the circuit branches have the same value when the circuit is in resonance and when detuned; i.e., $I_L \approx I_C \approx \frac{2}{200} = 0.01 \text{ a} = 10 \text{ ma}$. However, if the generator develops 200 volts and its internal resistance is equal to 1,000,000 ohms, the supply current consumed by the tuned circuit will be given by the following expression: $I \approx \frac{200}{1{,}000{,}000} = 0.0002 \text{ a} = 0.2 \text{ ma}$, voltage across the tuned circuit at resonance will be given by: $U = 0.0002 \times 10{,}000 = 2$ volts, while the currents

through the branches of the circuit will still remain at the same value of $I_L = I_C = 10$ ma. However, as soon as the circuit is detuned, the voltage and the currents in the circuit branches are decreased 10 times:

$$U = 0.0002 \times 1{,}000 = 0.2 \text{ v},$$

$$I_L = I_C = \frac{0.2}{200} = 0.001 \text{ a} = 1 \text{ ma}.$$

The derivation of the formula for the resonant impedance R_e is of interest. As mentioned above, power supplied by the generator is equal to power lost in the ohmic resistance of the tuned circuit,

$$\frac{U^2}{R_e} = I_{oc}^2 r.$$

Employing Ohm's law, substitute $\frac{U}{\varrho}$ for I_{oc}.

This will give the following:

$$\frac{U^2}{R_e} = \frac{U^2}{\varrho^2} r.$$

Dividing both sides of the equation by U^2, we have:

$$\frac{1}{R_e} = \frac{r}{\varrho^2}.$$

Hence:

$$R_e = \frac{\varrho^2}{r} = Q\varrho.$$

11. BANDWIDTH OF A TUNED CIRCUIT

During transmission of any type of signal by radio, the high-frequency aerial current of the radio transmitter may be represented by a combination of several currents, each of which has its own frequency (more detailed information is given in Chapter VIII). Electromagnetic waves radiated by the transmitting aerial have a similar complex character, and so do the currents generated by them in the receiving aerial.

In each type of radio transmission (radio telephony, radio telegraphy, television, etc.) the frequencies of these currents occupy a definite bandwidth. The frequency band on medium-wave broadcast transmission occupies about 9 kc. This signifies that a broadcast transmitter generates a complex current consisting of several currents, and the highest frequency is 9 kc higher than the lowest transmitted. For instance, for a broadcasting transmitter operating on a frequency of 173 kc ($\lambda = 1{,}734$ metres) the frequencies mentioned above will occupy a band from 168.5 to 177.5 kc. The frequency band of a commercial radio telephone transmitter usually does not take more than 2-2.5 kc of the spectrum, while in a radio telegraph transmitter it is much narrower still. On the other hand, such transmissions as television broadcasts take up bandwidths running into several megacycles.

Thus tuned circuits used by radio equipment have to produce several different frequencies within the limits of the bandwidth characterising a given type of transmission.

When a tuned circuit is being acted upon by electromotive forces of various frequencies, the strongest oscillations are obtained in the circuit when one of such forces has a frequency equal to the resonant frequency of the circuit, or close to it. When the frequency of the external e.m.f. differs by a considerable margin from the resonant frequency of the tuned circuit, i.e., when the latter is detuned in respect to the frequency of the external e.m.f., the amplitude of oscillations becomes comparatively small.

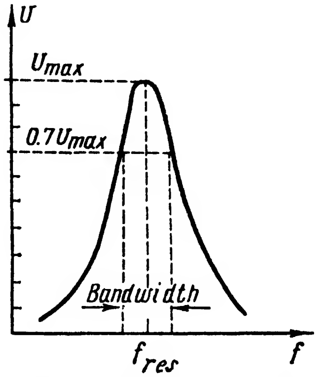


Fig. 14. Bandwidth of a tuned circuit

It may be said that each tuned circuit readily passes oscillations within a certain band of frequencies, located to both sides of the resonant frequency. Such a band is called the *bandwidth of a tuned circuit* and is conventionally determined in accordance with the resonance curve, at the level of 0.7 of the maximum voltage or current value corresponding to the resonant frequency (Fig. 14). In other words, it is considered that a tuned circuit readily passes oscillatory currents only when their amplitude does not drop by more than 30% of the resonant amplitude.

The bandwidth of a tuned circuit is sometimes called *the width of the resonance curve*. Fig. 10 has already been used to illustrate how the tuned circuit Q affects the shape of the resonance curve. As shown, the lower the quality factor of a tuned circuit, the broader band of frequencies will be passed by the given circuit. A higher resonance frequency of a tuned circuit also gives a broader bandwidth.

The dependence of a tuned circuit bandwidth upon the damping or Q of a given circuit is given by the following simple relation:

$$Bandwidth = \delta f_0 = \frac{f_0}{Q}.$$

Taking as an example a tuned circuit adjusted to a frequency of $f_0 = 2{,}000$ kc and possessing a damping value of $\delta = 0.01$, we can determine the bandwidth of the given circuit as equal to $0.01 \times \times 2{,}000 = 20$ kc.

It becomes clear that if we are to obtain a narrow bandwidth, we have to use tuned circuits with a high value of Q. Conversely, a tuned circuit designed for a broad bandwidth must either have a low value of Q or else must operate on a very high resonance frequency.

The formula given above clearly indicates that $f_0 = Q$ BW, where BW stands for the bandwidth. Since a tuned circuit of average quality has Q of at least 20, it becomes apparent that the operating frequency must be greater than the bandwidth by at least 20 times. For instance, television transmissions, in which the bandwidth is expressed in several megacycles, have to employ carrier frequencies of at least several dozens of megacycles, i.e., ultra-short waves.

It is always desirable to use tuned circuits with bandwidths corresponding to the frequency band assigned to a given type of transmission. If the bandwidth is narrower, distortion will occur due to poor passage of certain frequencies. Too broad a band is also undesirable, for it can result in interference from other radio stations operating on nearby frequencies.

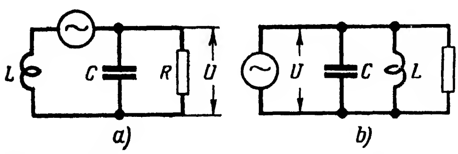


Fig. 15. Shunting of a tuned circuit by an ohmic resistance

High Q of a tuned circuit is not always advantageous. If a broad bandwidth is required, it is often found necessary to use tuned circuits with low Q. Q of a tuned circuit is lowered and the bandwidth is broadened when some ohmic resistance R, usually called a shunting resistance, is connected across the circuit (Fig. 15).

Studying this case, we see alternating voltage U, developed across the tuned circuit, applied to resistance R and setting a current flow through it. Naturally this resistance will consume a certain amount of power, given by Ohm's law as $\frac{U^2}{R}$. The smaller the value of resistance R, the greater will be the power losses in it and the larger will become the damping of the tuned circuit. If the value of resistance R is very low, it can be said that it will short-circuit one of the elements of the tuned circuit (the capacitor in Fig. 15*a*) or the whole tuned circuit (Fig. 15*b*). When this happens, the tuned circuit will not be able to function at all as an oscillatory system.

Radio circuits sometimes employ shunting resistors across tuned circuits for the purpose of broadening the bandwidth. In most cases, however, the shunting effect occurs as a result of connecting the tuned circuit with other components and circuits. Such shunting usually causes undesirable impairment of tuned circuit quality.

Internal resistance of the generator feeding a parallel-resonant circuit also influences Q of the tuned circuit and its bandwidth. This is easily understood from the following.

Assume that the generator stops functioning at a certain moment. As a result, the oscillations in the circuit will begin to damp down, while the internal resistance of the generator, still connected to the tuned circuit, will act as a shunting resistor increasing the damping effect.

The higher the internal resistance of the generator, the smaller will be its effect and, therefore, the sharper will be the resonance curve and the narrower the bandwidth, i.e., the more pronounced will be the resonant properties of the tuned circuit. At low values of R_i of the generator, Q of the tuned circuit is lowered to such a degree and the bandwidth becomes so broad that the tuned circuit practically loses its resonant properties.

We arrived at a similar conclusion concerning the effect of generator R_i upon the operation of a parallel-resonant circuit when studying the properties of this type of oscillatory circuit.

12. COUPLED CIRCUITS

When oscillatory energy is transferred from one circuit to another, such circuits are called coupled circuits.

In other words, *tuned circuits are said to be coupled when oscillations taking place in one of them are passed on to the other, setting up an oscillatory process in it.*

The greater the energy transferred from one tuned circuit to another, i.e., the stronger the influence exerted by one circuit upon the other, the closer is the coupling between them.

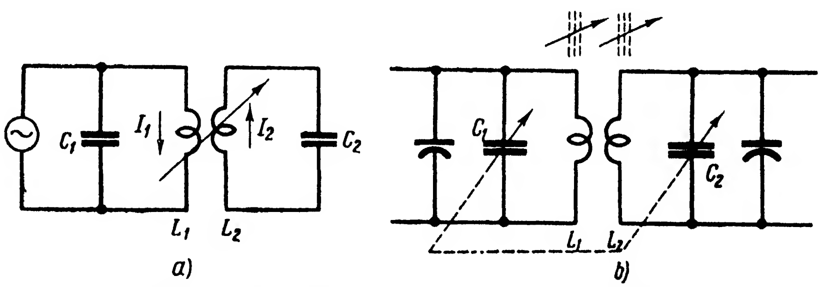


Fig. 16. Two inductively-coupled tuned circuits

The value of coupling is called the *coupling coefficient* C_{coup}, which can have a value from 0 to 1, or expressed in per cent — from 0 to 100%. When there is no coupling between the two tuned circuits, $C_{coup} = 0$. In radio circuits the value of the coupling coefficient can be anywhere between a fraction of one per cent to several per cent, in rare cases reaching several dozen per cent. A 100% value of the coupling coefficient is hardly ever encountered.

There are several types of coupling.

Inductive or transformer coupling is the type most frequently employed. It is formed by mutual inductance between the coils of tuned circuits. A schematic diagram of inductive coupling is shown in Fig. 16.

Circuit L_1C_1, *to which the generator energy is fed, is called the primary circuit. Circuit* L_2C_2, *receiving the energy from the primary* circuit, is called the *secondary circuit.*

The operating principle of inductive coupling is as follows. Current I_1, flowing through coil L_1 of the primary circuit, sets up a magnetic field around the coil. Lines of force of the given field interlink with the turns of coil L_2 and induce e.m.f. in it, this e.m.f. causing a current flow I_2 in the secondary circuit. Thus in inductively-coupled circuits the energy is transferred from one circuit to the other via the magnetic flux: Any transformer is an example of inductive coupling. The two coils which inductively couple high-frequency circuits constitute a *high-frequency transformer.*

Inductive coupling can be fixed or variable. Constructionally, fixed inductive coupling is represented by two single-layer or multi-layer coils, usually wound side by side on the same coil form. Variable inductive coupling is obtained by changing the distance between the two coils or by changing their relative position. In circuit diagrams variable inductive coupling is indicated by an arrow crossing the coupled coils, as shown in Fig. 16*a*.

The physical meaning of the coupling coefficient can be readily grasped by studying a case of inductive coupling. If inductances L_1 and L_2 are equal and there are no other coils in the tuned circuits, the coupling coefficient shows what part of the full magnetic flux Φ_1 of coil L_1 is taken up by magnetic flux Φ_c, the latter interlinking the two coils and coupling the circuits. For instance, if Φ_c is equal to 20% of Φ_1, the coupling coefficient is 0.2.

Circuits are adjusted to resonance in order to obtain maximum values of current and voltage. Either parallel or series resonance takes place in the primary tuned circuit, depending upon the connection of the circuit to the generator.

An inductive coupling series resonance is usually obtained in the secondary circuit.

This is because coil L_2 acts as a generator in the secondary circuit. This coil is series-connected with the rest of the tuned circuit; hence, this is a case of series resonance.

In practice circuits are adjusted to resonance to obtain maximum current in the secondary tuned circuit, the order of adjustment being the following. First, the primary circuit is tuned to resonance with the generator to obtain the maximum current value in this circuit, and only then the secondary tuned circuit is tuned to resonance with the primary circuit. After the secondary circuit has been adjusted, the primary circuit must be slightly retuned because the secondary circuit, coming into resonance with the primary one, upsets the resonant condition of the latter. Generally speaking, any change in adjustment of one of the tuned circuits slightly detunes the other coupled circuit. This is why each circuit must be additionally retuned (trimmed) to restore the condition of resonance.

To facilitate the tuning of two oscillatory circuits coupled to each other at a fixed value of coupling coefficient, their variable capacitors are frequently united (ganged) by placing both rotors on a common tuning shaft, which may be rotated by a single tuning control. In circuit diagrams capacitor ganging is indicated by a dotted line connecting the ganged capacitors (Fig. 16*b*).

Tuned circuits capacitances are equalised by means of small trimming capacitors of semi-variable type, whose capacitance can be adjusted within certain limits. These trimming capacitors are connected in parallel with the main capacitors (Fig. 16*b*).

Coil inductances are equalised by adjusting the position of cores inserted in the coils and made of some magnetic dielectric (carbonyl iron, alsifer, ferrite, etc.). Schematic representation of the cores is given in Fig. 16*b*, and these components are dealt with in a greater detail in Section 14.

In studying the operation of coupled circuits it is necessary to take into consideration the reverse influence exerted by the secondary circuit upon the primary. Current I_2, set up in the secondary circuit, flows through coil L_2 and creates magnetic flux around this coil. Some portion of this flux interlinks with the turns of coil L_1 and induces a certain value of e.m.f. in it. This e.m.f. acts in the opposite direction to the primary current I_1, reducing its value. In other words, the secondary circuit introduces or reflects some additional impedance into the primary circuit, and this impedance is, accordingly, known as *reflected impedance*.

When the secondary circuit is resonant with the generator frequency, it reflects into the primary circuit only ohmic resistance. The closer the coupling, the larger is the value of this resistance. Its value characterises the transfer of a certain amount of energy from the primary circuit to the secondary. When the secondary circuit is not adjusted to exact resonance with the generator frequency, it reflects into the primary circuit both ohmic resistance and a certain amount of reactance, which may be either inductive or capacitive, depending upon the direction in which the secondary circuit has been detuned. Thus a detuned secondary circuit upsets the tuning of the primary circuit with which it is coupled.

A curve showing the dependence of the current or voltage of a secondary circuit upon the frequency of the generator is also the resonance curve for a system of two coupled circuits. The shape of such curve depends upon the value of coupling. The looser the coupling, the sharper the resonance (Fig. 17).

If the coupling is gradually made closer, the resonance curve will flatten out at the top until, at a certain value of coupling, it will acquire a double-humped shape. The value of coupling at which the resonance curve shape is changed from the single-humped variety to the double-humped one is known as *the critical coupling*.

On critical coupling (dealing with tuned circuits of similar design) current, voltage and power of oscillations in the secondary circuit are at their maximum values, compared with those obtained on looser or closer coupling. The critical coupling is therefore referred to as *the optimum coupling,* i.e., the most advantageous value of coupling.*

It should be borne in mind that the optimum coupling is most advantageous only from the point of obtaining maximum power in the secondary circuit.

In practice the value of the optimum coupling coefficient in tuned circuits of similar design is equal to the damping value of each tuned circuit. Thus, for instance, if each one of the tuned circuits has a δ value of 0.02 (or $Q = 50$), optimum coupling will be obtained when the coupling coefficient is equal to 0.02 or 2%.

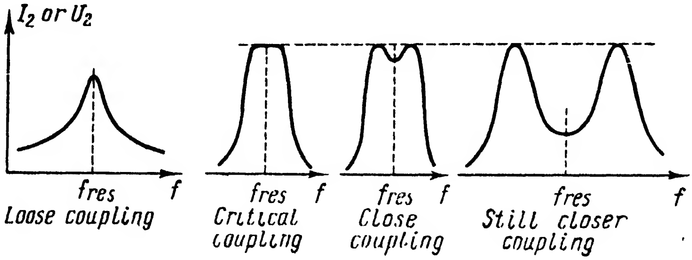


Fig. 17. Resonance curves for two coupled tuned circuits at different values of coupling

When the degree of coupling is less than the critical value, the coupling is called *loose*. On loose coupling the resonance curve has nearly the same shape as in the case of a single (uncoupled) circuit. When the coupling value is greater than the critical, it is called *close* coupling. If the coupling value is increased above the critical value, the depression in the resonance curve becomes more pronounced and the frequency difference of the two humps on the given curve increases (Fig. 17).

Critical or close coupling (when the depression between the humps is small) considerably broadens the bandwidth and is therefore used in certain radio receiver circuits (see Chapter IX). On close coupling, energy transfer from the primary to the secondary circuit takes place at high efficiency (over 50%), i.e., the power in the secondary tuned circuit is greater than that lost in the primary tuned circuit. It is because of this that close coupling is used in high-power circuits, such as those employed in radio transmitters. Loose cou-

* The values of critical and optimum coupling are somewhat at variance for tuned circuits with different parameters.

pling is employed when it is not required to transfer maximum energy from the primary circuit to the secondary or when high efficiency is not sought, but when it is important that the secondary circuit exerts only slight influence upon the primary. This type of coupling is frequently used in radio measurements.

Capacitive coupling employs a coupling capacitor (Fig. 18). Here energy is transferred from the primary circuit to the secondary through the electrical field. Capacitive coupling often exists between various parts of a circuit where it is not desired and impairs normal operation of the circuit. In such cases it is called parasitic, and measures are taken towards its elimination or the lessening of its effects. The circuit shown in Fig. 18*a*, where the coupling capacitor C_c is not included in the primary or secondary circuits, is called *external capacitive coupling*. Fig. 18*b* depicts an *internal capacitive coupling* where the coupling capacitor is connected simultaneously in the primary and secondary circuits, in series with each capacitor C_1

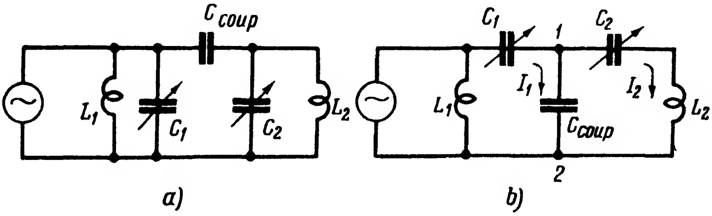


Fig. 18. Capacitive coupling circuits: *a)* external capacitive coupling; *b)* internal capacitive coupling

and C_2 of these two circuits. If it is desired to adjust the degree of coupling, the coupling capacitor must be made variable.

In a circuit with external capacitive coupling the primary tuned circuit voltage is applied through the coupling capacitor to the secondary tuned circuit and causes a current flow through it. The larger the capacity of the coupling capacitor, the less opposition it presents to the alternating current and the closer is the coupling, i.e., the greater is the energy transferred to the tuned circuit L_2C_2. In practical circuits loose coupling is obtained by using small coupling capacitors with a capacitance of only several picofarads (which is much smaller than the capacitance of C_1 and C_2). Parasitic coupling between various components and parts of a circuit arises owing to the external type of capacitive coupling, as in this case a very small capacitance is sufficient to form a parasitic coupling. Inductive coupling is always accompanied by a certain amount of external capacitive coupling due to the capacitance existing between the coils and connecting leads.

In a circuit with internal capacitive coupling (Fig. 18*b*), voltage created across the coupling capacitor as a result of current flow I_1

through it acts upon the secondary tuned circuit C_2L_2, producing current I_2 in it. It may be stated that at point *1* (or *2*) the current is *branched off and a part of it enters the secondary circuit. In direct* opposition to what takes place in Fig. 18*a*, we have here to reduce the value of the coupling capacitor to increase the degree of coupling. In this case the reactance presented by the coupling capacitor to current I_1 will be increased, which will also increase the voltage drop across this capacitor, the effect of which upon the secondary tuned circuit will be an increase of current I_2. To decrease the degree of coupling in this type of circuit, the value of the coupling capacitor is made large — thousands and even tens of thousands of picofarads (much higher than the values of C_1 and C_2).

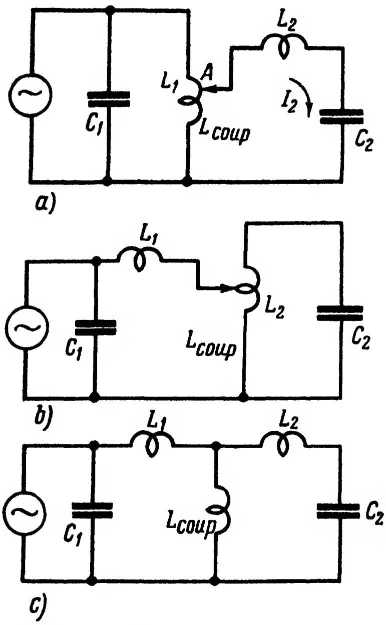


Fig. 19. Various autotransformer coupling arrangements

Autotransformer coupling differs from other types of coupling arrangements in that both tuned circuits use a common coil. The energy is transferred partly through the magnetic field and partly through the direct electrical connection between the tuned circuits.

In a circuit of this type, shown in Fig. 19*a*, coil L_1 is a part of the primary circuit, and a part of this coil (L_c) belongs to the secondary circuit. The voltage developed across L_c causes current I_2 to flow. Coil L_1 acts as a step-down transformer or as an inductive voltage divider. Additional coil L_2, together with L_c, forms the inductance of the secondary circuit. The larger the part of coil L_c in both tuned circuits, the closer is the coupling. In the arrangement shown in Fig. 19*b* coil L_2 is a part of the secondary tuned circuit and acts as a step-up autotransformer. Part L_c of coil L_2 is included in the primary tuned circuit, in which an additional coil L_1 is included. In this type of circuit the degree of coupling also increases with an increase of L_c.

In the arrangement shown in Fig. 19*c* the smaller the value of L_c in comparison with L_1 and L_2, the looser is the coupling between the tuned circuits.

When the degree of coupling is fixed, wire tapped to coil L_c is secured permanently to it. Variable coupling requires either a switch to select different numbers of turns, or—when the coil is wound with bare wire—a slider which may be moved along the winding

and set to any required position. Schematically, such a slider is indicated by an arrow.

Combined coupling is occasionally employed and is represented by two types of coupling, usually inductive and capacitive. As an example, Fig. 20*a* shows an arrangement with external capacitive coupling, while Fig. 20*b*—an arrangement with internal capacitive coupling.

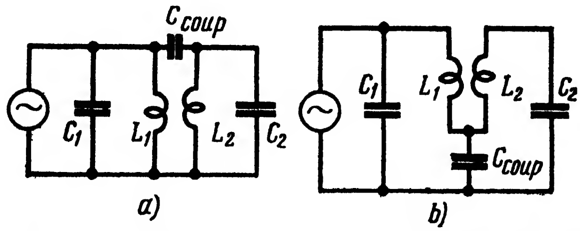


Fig. 20. A circuit of combined inductive-capacitive coupling

The various types of coupling studied above can also be used by both usual alternating current circuits or by oscillatory and aperiodic circuits.

13. SHIELDING

It often becomes necessary to take measures towards the elimination of undesirable capacitive or inductive coupling between tuned circuits or components in radio equipment. One of such measures consists in moving the offending tuned circuits or wires away from each other. This, however, is not a very good solution, as it entails an increase in the overall dimensions of the equipment. *Shielding* offers a better alternative.

By shielding is meant the protection of a tuned circuit or its components from the influence of another tuned circuit, by means of metal sheets, known as shields.

The shields are used to eliminate inductive as well as capacitive types of coupling. On low frequencies elimination of inductive coupling calls for shields made of ferromagnetic materials, i.e., from 0.5 to 1.5 mm sheet steel. In this case use is made of the effect of drawing in of magnetic lines of force by the steel. Owing to the high magnetic permeability of the steel, the magnetic lines of force are made to complete their circuit in the shield and do not leave the artificial boundaries thus established.

On high frequencies the best type of shielding is provided by diamagnetic high-conductance metals. Preference is usually given to 0.3-1 mm sheet copper or aluminium. The action of such shields is as follows. The magnetic flux of a coil induces currents in the

shield, and according to Lenz's law, these currents set up their own magnetic flux. This flux, being opposite in direction to the *offending flux, almost completely neutralises the latter outside the shield boundaries. To keep down the losses associated with the* creation of currents in the shield, the latter should not be placed too near to the coils of tuned circuits. It is also desirable that the diameter and length of shields are correspondingly not less than the doubled value of coil winding diameter and length. If this is observed the quality of a tuned circuit is not materially impaired by the shielding. It should be kept in mind that a shield considerably decreases the inductance of a coil (by 10-20%).

Elimination of parasitic capacitive coupling also requires shields made of high-conductance metals and connected with the radio equipment chassis. The operating principle of the shields is easily understood from Fig. 21.

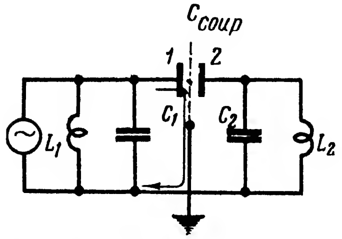


Fig. 21. Elimination of parasitic capacitive coupling with the help of a shield

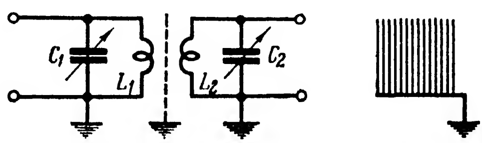


Fig. 22. Application of a shield for elimination of parasitic capacitive coupling between inductively-coupled tuned circuits

Two tuned circuits L_1C_1 and L_2C_2 are linked by a parasitic capacitive coupling denoted as capacitor C_c. The common wire of the two circuits is usually connected to the chassis (earth). If an earthed shield is placed between the two capacitor plates *1* and *2* (which are, in reality, represented by wires or components of the tuned circuit), current from the tuned circuit L_1C_1 will flow through the capacitance between plate *1* and the shield, returning to the primary tuned circuit and not entering tuned circuit L_2C_2.

Schematically the shields are indicated by broken lines.

It sometimes becomes necessary to eliminate parasitic capacitive coupling between two coils in order to obtain a pure inductive coupling. A solid metal shield is not suitable in this case because it eliminates both types of coupling. The problem is solved by application of an electrostatic shield, which eliminates capacitive coupling only. A shield of this kind consists of a row of thin wires connected to one another and earthed at one end. No induced currents appear in this type of shield, as no closed circuits are offered by the shield to such currents. Fig. 22 shows a circuit representation of an electrostatic shield and the principle of design of one of its versions.

14. TYPES OF TUNED CIRCUITS AND THEIR COMPONENTS

Adjustment of Tuned Circuits

In radio transmitters and receivers the tuned circuits are usually designed for a certain frequency range and differ in respect to frequency changing methods, i.e., methods of tuning to various frequencies. The most common type of tuned circuit employs a fixed inductance and a variable capacitance. Such circuits have already been studied above. Fig. 23*a* illustrates another type of tuned circuit, where the capacitance is fixed and inductance is varied (see more information on variometers below).

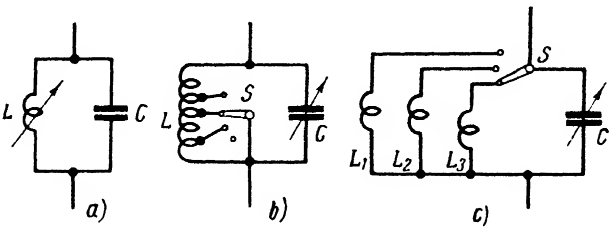


Fig. 23. Types of tuned circuits for narrow and broad frequency coverage

Such versions of tuned circuits can give frequency or wavelength change (coverage) not in excess of 2 or 3.

To broaden the coverage, a method of changing both the capacitance and inductance is employed, as illustrated in Fig. 23*b* and *c*. Coarse frequency change in the tuned circuit shown in Fig. 23*b* is secured by means of switch *S*, changing the number of coil winding turns step-by-step. Fine tuning is secured by means of a variable capacitor. The circuit illustrated in Fig. 23*c* has a separate coil for each frequency band, i.e., for each part of the frequency range, split into bands.

All tuned circuits employing tapped coils have the inherent disadvantage that the unused part of the turns, short-circuited by switch *S*, is coupled to the working part of the coil and absorbs a certain amount of high-frequency energy from the circuit, thereby impairing the circuit quality. If the unused part of the coil were not shorted by the switch, it would, because of its own capacitance, constitute a parasitic tuned circuit. The latter is always tuned to a certain frequency, on which it will rob the operating tuned circuit of high-frequency energy to a particularly noticeable degree. This effect can be prevented in a tuned circuit where there is a separate coil for each frequency band, and placing each coil in a separate shield will solve the problem.

Other types of tuned circuits are sometimes encountered in radio equipment. Thus *some transmitters switch capacitors of different capacitance into the tuned circuits* when switching over from one band to another.

Coils with Fixed Inductance

When a coil requires a comparatively small number of turns it is wound with a single layer of insulated or bare wire on a coil form made of cardboard, porcelain, or some other insulating material (Fig. 24*a*, *b*). The insulated wire usually has a thickness of 0.2-1 mm. An improved design, providing better inductance stability, employs a copper or silver spiral embedded into the surface of a ceramic coil form.

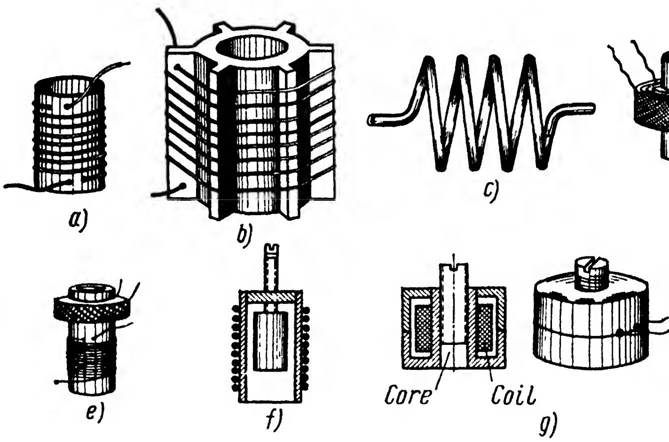


Fig. 24. Types of coils employed in tuned circuits

As previously explained, high-frequency currents flow only through the surface layer of a conductor. Hence, the ohmic resistance presented by the conductor to high-frequency currents can be lowered if the conductor surface is increased. For this purpose the so-called *litz wire* is sometimes used on long, medium and even certain short waves. A litz wire consists of a large number of thin enamelled strands, insulated one from another. The total surface of the strands in a litz wire is considerably larger than the surface of a solid conductor of comparative overall diameter. Distributed capacity of a coil is frequently undesirable and is reduced by winding the turns not side-by-side but at some distance from each other. Coils used by tuned circuits of transmitters, and particularly short-wave transmitters, are usually made of thick bare wire or even copper tubing. In this case, no coil form is required to support the winding (Fig. 24*c*).

Coils requiring windings with a large number of turns are usually multi-layer type coils. Such coils most frequently use "Universal" (zigzag) windings, in which the wire passes from one edge of the coil to the other (Fig. 24*d*). When a coil consists of several sections, spaces are provided between them (Fig. 24*e*) to lessen the influence of the unused turns and to decrease the distributed capacity of the coil. In such coils the long-wave winding is of multi-layer type, while the shorter-wave winding is wound in a single layer (Fig. 24*e*). The coils are protected from moisture by coating them with varnish or by impregnating them with paraffin or some other insulating substance. Such coating and impregnation, however, increase the losses and impair the coil Q, particularly when poor-quality varnishes are used.

Magnetic dielectrics have been successfully employed in the manufacture of coil cores for high-frequency tuned circuits. Such materials are usually made of pressed iron powder (or powder of other magnetic materials). The powder particles are glued together by some insulating substance, such as varnish. The magnetic dielectric materials most frequently used are carbonyl iron, powdered permalloy, ferrite, alsifer and sometimes magnetite. Power losses in the cores made of magnetic dielectric materials are very low even on high frequencies. Such cores increase the inductance of a coil several times and make it possible to obtain required values of inductance with a considerably smaller number of turns, and hence much shorter lengths of wire. This reduces the loss resistance and improves the Q of the tuned circuit.

Cores made of magnetic dielectric materials have various shapes. The simplest is a cylindrical core located inside the coil (Fig. 24*f*). Insertion of the core farther into the coil increases the coil inductance. Greater inductance changes are secured by a closed core shaped as a cylindrical container (Fig. 24*g*), consisting of two halves and surrounding the coil from all sides. The central part of the core may be screwed out, which will somewhat decrease the inductance. Coils with this type of core need no shields because the magnetic flux is concentrated in the core and is not dispersed in the air.

On long and medium waves the Q of cores made of magnetic dielectric materials reaches 400-500 (in air-core coils it is difficult to obtain Q values over 200). On short and ultra-short waves magnetic dielectric cores give a smaller gain in Q values. On these waves small cores are used not for the purpose of improving the Q, but for inductance adjustments.

In air-core coils a limited inductance adjustment ("coil trimming") is performed by other methods. For instance, part of the coil winding turns is wound separately, and moving these turns to change their position (in relation to the rest of the winding) changes the inductance. Multi-layer coils are sometimes made of two similar sections, and the coil is trimmed by varying the distance between them.

Sometimes a short-circuited wire turn or a diamagnetic (copper, brass, aluminium) *disc or rod is placed inside the coil. Currents are induced in such wire turn or a piece of metal, and the magnetic* field created by them weakens that of the coil and decreases its inductance. Such a trimming element must be shifted to adjust the coil inductance to the required value. When the adjustment is completed, the trimming element is fixed in the proper position to avoid accidental changes of inductance.

Recently printed circuits have often been employed in the manufacture of radio equipment, particularly its portable varieties. In a printed circuit various components and wiring are represented by thin conducting strips laid in various ways over the surface of an insulated panel. In such circuits the coils generally have the shape of metal spirals placed over the insulating surface.

Coils with Variable Inductance

Coils whose inductance can be adjusted within wide limits are used for continuous tuning of circuits within a wavelength range and are sometimes referred to as variometers. Fig. 25*a* shows the design and schematic representation of a variometer with a moving coil, which consists of two coils usually connected in series with each other. The external coil is stationary and is called the stator, while the inner coil — the rotor — is fixed on a shaft and can be rotated. When the rotor is turned, the total inductance of the variometer changes due to the change of mutual inductance between the coils. If the two coils are placed at right angles to each other (position *2* in Fig. 25*b*), their magnetic fields do not interact. In this position the mutual inductance is zero

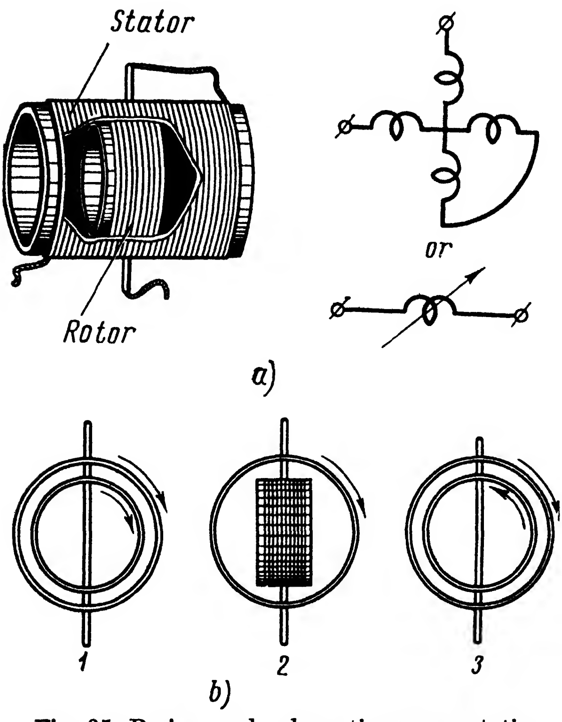


Fig. 25. Design and schematic representation of a moving-coil variometer

and the total inductance L is equal to the sum of inductance of the coils. Turning the rotor through 90° in either direction places the coils in relative position *1* or *3* (Fig. 25*b*). In this case the coil axes coincide. However, in position *1* currents flow through the coils in the same direction, the magnetic field is increased and the total inductance becomes larger. In position *3* the inductance is lowered because the currents through the coils flow in opposite directions and the magnetic field is decreased. Thus turning the rotor through 180° gives a continuous change of inductance from some minimum to maximum value.

Maximum possible inductance change requires that the rotor turns come as close as possible to the stator winding. However, even in the best designs the maximum inductance change does not exceed 8 times. Some designers employ switching the coil connection from series to parallel, which decreases inductance. A general disadvantage of all moving-coil variometers is that the whole length of the wire of both coils remains connected in the tuned circuit when the rotor is set for minimum inductance.

Changing the coil inductance within wide limits can be achieved by continuous changing of the number of coil turns. This was formerly accomplished by employing coils made like rheostats, whose sliders moved along the winding, changing the inductance turn-by-turn; but the slider was notorious for poor contact and, besides, short-circuited adjacent turns.

Ferro variometers, in which the coil inductance is varied by a shifting magnetic dielectric core, are an improvement over the old-type variometers. Depending upon the design of the core, inductance changes as high as 5-10 times have been optained.

In some cases continuous change of inductance is obtained by shifting the position of a piece of diamagnetic material (shaped as a disc, flap, cylinder or ring) placed in the magnetic field of the coil.

High-Frequency Transformers

There are various types of high-frequency transformers which employ the principle of inductive coupling. Single-layer coils are frequently wound side by side on the same coil form or are inserted in each other (Fig. 26*a* and *b*). This gives a coupling coefficient not in excess of 0.5. Closer coupling may be obtained by winding one coil between two halves of winding belonging to the other one (Fig. 26*c*). A still higher value of coupling coefficient (up to 0.8) is obtained in "unity coupling" when, during the process of winding the coils, the turns of one coil are interlaced with the turns of the other one (Fig. 26*d*).

High-frequency transformers with two multi-layer windings (Fig. 26*f*) have found a wide application. Coupling coefficients of

up to 0.8 have been reached with these units. The coils of high-frequency transformers are often supplied with cores made of magnetic dielectric materials. Placing such cores in the space between the windings considerably increases the coupling coefficient. The coils are moved in the transformer in relation to each other and, once the necessary coupling is obtained, they are locked in position.

Variable inductive coupling is sometimes arranged in a variometer-like device shown in Fig. 25. In such a circuit the variometer rotor winding is connected to one of the coupled circuits, and the

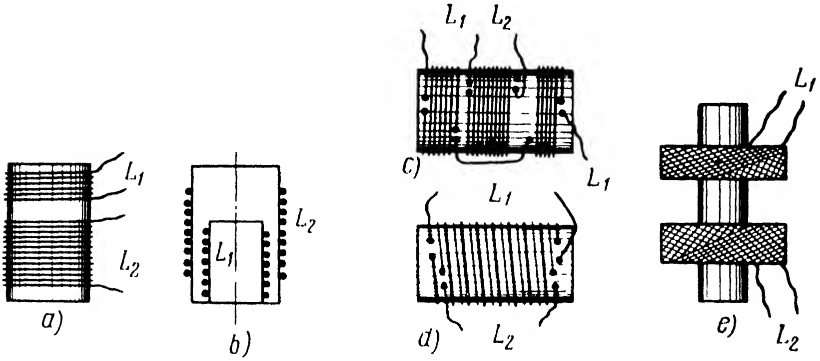


Fig 26. Types of high-frequency transformers

stator winding to the other. Turning the rotor through 90°, e.g., from position *1* to position *2* (Fig. 25), changes the coupling from its greatest value to its smallest.

Fixed Capacitors

The basic parameters of any capacitor are: *capacitance value, stability and accuracy of capacitance, working voltage* U_w, *test voltage* U_t, *losses, and insulation resistance.*

Capacitance is expressed in picofarads (pf) or microfarads (mfd).

Of great importance is accuracy of capacitance value and stability of capacitance, both of which depend upon the design of a capacitor. There are many causes for capacitance changes. Heating, that changes the geometric dimensions of a capacitor and dielectric permeability of its insulation, exerts a particularly strong influence. Capacitors with low accuracy and stability of capacitance are used only in cases where considerable capacitance changes do not materially affect the operation of a given circuit.

The accuracy of capacitance value is sometimes given directly on capacitors and is indicated in the following way: $C = 500\,\text{pf} \pm 10\%$, which means that the capacitance does not vary by more

than 10% (50 pf) in either direction from its rated value and is always within the limits of 450 to 550 pf.

The working voltage is that maximum voltage at which the capacitor can safely operate over long periods of time. The value of U_w must not be less than the peak value of alternating voltage applied to the capacitor. When the capacitor is used with pulsating currents, its working voltage must not be less than the maximum voltage value equal to the sum of the direct current component and the peak value of the alternating current component.

Leakage resistance is the value of leakage current appearing due to dielectric imperfections. The value of such resistance of a capacitor is usually expressed in hundreds and thousands of megohms.

The quality of a capacitor is also determined by losses in power occurring when alternating current is passed through it. In a good capacitor these losses are negligible and are much smaller than the losses occurring in tuned circuit coils.

Described below are various types of fixed capacitors used in oscillatory circuits.

Air capacitors, i.e., capacitors using air as dielectric, have capacitances not exceeding several hundred picofarads. These capacitors are usually made of aluminium plates secured by small bolts and washers, the latter providing the air gap between the metallic plates, the value of which is determined by working voltage of the capacitor. Capacitors of this type are mounted on plates made of ceramic or other insulating material. Air capacitors have the smallest losses, as compared with all other types of capacitors. However, air capacitors occupy considerable space in radio equipment and are rather expensive. They are used in the tuned circuits of transmitters.

Ceramic capacitors. These are only slightly inferior in quality to air capacitors, but they are inexpensive and are of small dimensions. Special ceramic materials characterised by low losses at high frequencies are used as dielectrics in these capacitors. The most commonly used are radio porcelain, tycond, ultrasteatite, tetrabar, tydol, etc.

In ceramic capacitors the plates consist of layers of silver spread over ceramic surface and usually coated with varnish.

The capacitance values of these capacitors range from several picofarads up to hundreds of picofarads, with working voltages from several hundreds to several thousands of volts. Ceramic capacitors possess high stability, low losses and high insulation resistance. They are manufactured in a variety of types, e.g., disc, tubular, barrel, and pot-type (Fig. 27).

Of special interest are tycond capacitors, whose capacitance decreases during heating, instead of increasing as it does in capacitors of other types. Hence tycond capacitors may be used to compensate thermal capacitance changes, thus improving the frequency stability of tuned circuits.

Mica capacitors. **Before the invention of ceramic capacitors, mica capacitors were widely used in tuned circuits. They have now been largely replaced by the superior ceramic types. The capacitance** *range of mica capacitors extends from tens to tens of thousands of picofarads, the working voltages being from several hundreds to* several thousands of volts. The capacitor plates are made of lead-tin or aluminium foil. In mica capacitors of recent manufacture, layers of silver, placed on mica insulation, serve as capacitor plates. Mica capacitors are now usually of the potted type (i.e., they are manufactured sealed in plastics). Such a design protects them from moisture and mechanical damage.

Capacitors intended for operation in low-power equipment are grouped into three classes according to accuracy; the capacitance value tolerance is $\pm 5\%$ for the first class, $\pm 10\%$ for the second, and $\pm 20\%$ for the third. Capacitors with higher accuracy than $\pm 5\%$ are manufactured for special purposes. Several types of mica capacitors are shown in Fig. 27.

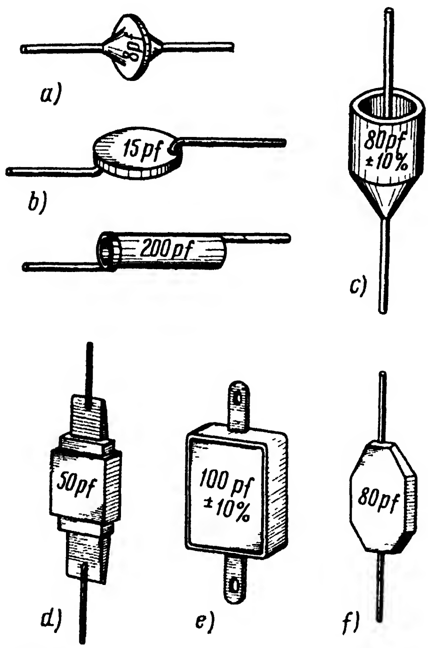


Fig. 27. Various types of capacitors. Ceramic capacitors: *a)* disc-type, *b)* tubular-type, *c)* pot-type. Mica capacitors: *d)* in metal plating, *e) and f)* moulded in plastics

Paper capacitors. As far as electrical quality is concerned, paper capacitors are much inferior to the mica capacitors. They should not be used in high-frequency circuits as the losses in paper insulation are excessive. Paper capacitors are applicable only in low-frequency circuits and in the by-pass service of some high-frequency circuits.

In those rare cases when oscillatory circuits employ paper capacitors, the latter should be of the non-inductive type. They are usually made of interlaced foil and paper strips tightly rolled into spirals. The ends of the foil strips serve as capacitor terminals.

Capacitors of this design possess some inductance, and their high-frequency reactance can be considerable. In non-inductive paper capacitors, foil strips are staggered, and after they are rolled into spirals their edges, located at the opposite sides of a capacitor, are clamped and connected to the terminal wires. The lead from the

external plate of such a capacitor is usually marked and should be connected to chassis (earth).

In capacitors designed for the printed circuits the plates are usually represented by metallic layers placed on both sides of insulated panels carrying printed circuit conductors.

Variable Capacitors

Oscillatory circuits requiring variable capacitance employ air-dielectric capacitors almost exclusively. Solid-dielectric variable capacitors are very seldom encountered.

When a variable air capacitor is connected into a tuned circuit its capacitance and the frequency of the tuned circuit can vary in different ways, depending upon the shape of the capacitor plates. Therefore, variable capacitors designed for respective types of services employ plates the movement of which gives straight-line change in any desired parameter (capacitance, frequency, wavelength), as desired. The capacitors are hence respectively called straight-line capacitance, straight-line frequency, straight-line wavelength capacitors. There is also a logarithmic capacitor, discussed below together with other types of variable capacitors.

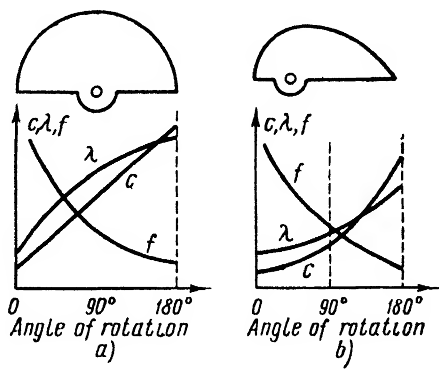


Fig. 28. Plate shapes and tuning curves of an oscillatory circuit using: *a)* a straight-capacitance capacitor, or *b)* logarithmic capacitor

Straight-line capacitance capacitor (with semicircular plates). Fig. 28*a* shows the shape of plates and changes of capacity C, frequency f and wavelength λ of the tuned circuit comprising the capacitor.

When the capacitor rotor is turned, the capacitance changes proportionally to the angle of rotation (the capacitance curve is a straight line), while the frequency changes as shown by respective curves in Fig. 28. At the beginning of rotation, i.e., when the capacitance is small, the frequency changes sharply (steep part of the curve), reducing its slope on further rotation. Uneven frequency change makes it inconvenient to use this type of capacitor for tuning and calibration in terms of frequency. Straight-line capacitance capacitors are chiefly used in measuring equipment.

Straight-line frequency capacitor, straight-line wavelength capacitor and logarithmic capacitor. All these are noted for the elongated shape of their plates, resembling the shape of a bird's wing and

slightly different in each of the three types. The frequency of a tuned circuit comprising a straight-line frequency capacitor changes evenly with rotation. The straight-line wavelength capacitor offers an even change of wavelength. The logarithmic capacitor, which is *most frequently employed in radio circuits, is supplied with plates of configuration shown in Fig. 28b. This drawing also gives the curves* for the change of capacitance, frequency and wavelength of a tuned circuit comprising such a capacitor. In logarithmic capacitors changes of wavelength and frequency, expressed in per cent, are constant over the entire tuning range. Thus, for example, if a capacitor setting gives 300 metres at some scale division and the wavelength changes by 1%, i.e., by 3 metres when the capacitor is rotated through one division, then a similar percentage change will occur when such capacitor is rotated through any other division of the scale. For instance, when the capacitor is set to a wavelength of 500 metres, further advance of the capacitor rotor over one scale division will change the wavelength by 5 metres (1%).

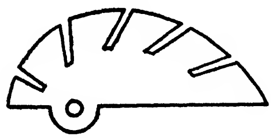


Fig. 29. Rotor plate with slits for capacitance matching

Modern receivers and transmitters use ganged capacitors consisting of two or three sections and driven by a common shaft, each one of such sections represented by an individual capacitor belonging to an independent tuned circuit. A common shaft carries the rotors of all the ganged capacitors. The rotors are frequently connected electrically to each other and to the capacitor frame, while the stators are isolated, each having an independent terminal.

The problem usually encountered in circuits employing ganged capacitors is that of capacitor tracking, i.e., obtaining an equal amount of capacitance change in each capacitor as the gang is rotated: When logarithmic capacitors are used this problem is best solved by equalising the minimum capacitance values of the tuned circuits by means of trimming capacitors. If similar tuning is obtained at the beginning of the scale (minimum capacitance) for all the ganged sections, then the tracking will be preserved at all other parts of the scale connected with the shaft. This, of course, requires that the capacities of the ganged sections change equally as the capacitor shaft is rotated. Unfortunately, it is rather difficult to manufacture variable capacitors which are exactly equal in their electrical characteristics. To match the capacitors with slightly different characteristics, the rotor plates of such capacitors are supplied with slits, as shown in Fig. 29. During circuit adjustment to obtain exact tracking of the ganged capacitors, various sections of rotor plates provided with such slits are slightly bent in or out until all the capacitors equally change their capacitance for an equal angle of rotation.

The capacitance value of usual variable capacitors does not

exceed several hundreds of picofarads, the working voltage of capacitors being determined by the distance between rotor and stator plates. In capacitors designed for service in low-voltage equipment (radio receivers), *the working voltage is several hundred volts. High-power transmitters use variable capacitors with working* voltage of several thousand volts. The minimum capacitance of most variable capacitors is generally 5-10% of their maximum capacitance.

Trimming capacitors. Trimming capacitors are mainly used for tracking adjustments in ganged capacitors. The "trimmers", as they are called, are connected across the variable capacitors of the gang. When the latter are set to their minimum capacitance, the trimmers are tuned to adjust each ganged section to similar resonant values.

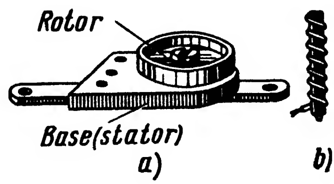


Fig. 30. Trimming capacitors: *a)* ceramic; *b)* "home-made" twisted-wire variety.

There are many different designs of trimming capacitors, one of the best being shown in Fig. 30*a*. In this design a ceramic disc serves as a dielectric and acts as rotor plate. The surface of the disc is covered with a layer of silver, shaped as a semicircle or a sector. The stationary base of the capacitor is also made of ceramic material carrying a silver coat and serving as stator.

Some trimmers incorporate a spring plate, which can be brought closer to a stationary plate or moved farther away from it by means of the tuning screw. In such designs a layer of mica is usually placed between the plates to prevent them from touching.

Some trimmers use a cylindrical construction, in which one side of the capacitor is a cylinder and the other is an insert, which moves in and out of the cylinder. Air or mica are used as dielectrics in such trimmers.

The concentric trimmer is supplied with cylinder-type stator and rotor, each one of which carries a set of concentric cylindrical ribs, the stator and rotor ribs meshing with each other on tuning, and having air dielectric between them.

Sometimes the trimmers are designed like miniature straight-line capacitance capacitors, carrying a small number of semicircular plates. Some designs use an arrangement in which a screw shifts a moving plate in relation to a stationary one.

In many cases the trimmer is designed for direct mounting on the chassis, so that either its rotor or stator is connected to the metal of the chassis (earth). Frequently trimming capacitors are installed directly on the frame of the main tuning capacitor.

"Home-made" capacitors, consisting of two insulated wires, are sometimes used by radio amateurs. The wires are either twisted together (without an actual electric contact between their conductors), *or else one of the wires is wrapped around the other (Fig. 30b).*

In the adjustment of a radio circuit, the trimming capacitors are usually locked in position once the required capacitance has been set. This is done by applying a dab of paint to the adjustment screws or rotor shafts of the trimmers.

Trimming capacitors are designed for maximum capacitance values from several picofarads to hundreds of picofarads.

15. SIMPLIFIED DESIGN OF TUNED CIRCUITS AND THEIR COMPONENTS

Thomson's formula is employed as the basis of tuned circuit design. For the sake of convenience, inductance is expressed in microhenries, capacitance in picofarads, and frequency in kilocycles; 2π is usually taken equal to 6.25, which gives an error of only 0.5%. In the final result an error of 5-10% is inevitable in any case, since it is impossible accurately to evaluate such factors as distributed capacitance of wiring, influence of coil shields and many other points.

The approximate formula for the calculation of the natural frequency of a tuned circuit can be written as follows:

$$f_{(\mathrm{kc})} = \frac{160{,}000}{\sqrt{L_{(\mathrm{microhenries})}C_{(\mathrm{picofarads})}}}.$$

In this formula C must be understood as the full capacitance of the tuned circuit, consisting of the capacitance of the tuning capacitor and the distributed capacitance of wiring. The latter usually calls for a 20-40 pf allowance.

Example 1. Find the wavelength and frequency range of a tuned circuit consisting of a 40-microhenry inductance coil and a variable capacitor with a coverage of 10-130 picofarads. Assume that the distributed capacitance of wiring is equal to 30 picofarads.

Solution.

The total capacitance of the tuned circuit is changeable from $C_{min} = 40$ pf to $C_{max} = 160$ pf, which gives an effective capacitance coverage of 4. Hence, λ and f coverage is equal to 2. From value of C_{min} we find the value of f_{max}, as follows:

$$f_{max} = \frac{160{,}000}{\sqrt{40\times 40}} = \frac{160{,}000}{40} = 4{,}000 \text{ kc}.$$

Hence $f_{min} = 2{,}000$ kc. The wavelength range, accordingly, extends from 75 to 150 metres.

In most practical cases, when the frequency or wavelength, as well as L or C, are known, the remaining parameters are determined with the help of the following formulas, where L is expressed in microhenries, C — in picofarads, and f — in kilocycles:

$$L = \frac{25\times 10^{9}}{f^{2}C}; \qquad C = \frac{25\times 10^{9}}{f^{2}L}.$$

Example 2. It is required to design a tuned circuit with a wavelength coverage from 30 to 60 metres (frequency coverage from 10,000 to 5,000 kc). The circuit will use a 20-220 pf capacitor. The distributed capacitance of wiring is 30 pf. Find the inductance of the coil.

Solution.

Total capacitance coverage of the tuned circuit is 5, because the capacitance changes from $C_{min} = 50$ pf to $C_{max} = 250$ pf. The wavelength and frequency coverage is $\sqrt{5}$, which is approximately equal to 2.2. We are seeking a wavelength coverage of 2 (from 30 to 60 metres) and apparently have a reserve of approximately 10%. It would be rational now to calculate the required inductance as follows, assuming C_{max} and f_{min}:

$$L = \frac{25 \times 10^9}{5{,}000^2 \times 250} = 4 \text{ microhenries.}$$

If it happens that, due to the inevitable calculation error, the frequency is somewhat lower (within 10%), the required range will still be secured. If, on the other hand, the frequency is somewhat higher than the needed value, we can easily lower it with a small parallel-connected capacitor. The inductivity could also be found on the basis of f_{max} and C_{min}, but in this case we might encounter a little difficulty; if the actual frequency were lower than the required value, we should have to reduce the inductivity or the capacitance of the tuned circuit, which is rather complex.

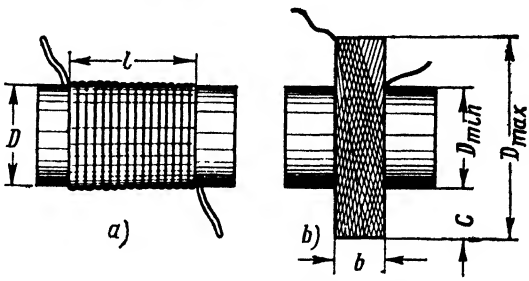


Fig. 31. Single-layer *(a)* and multi-layer *(b)* coils employed by tuned circuits.

When calculating the inductance of a single-layer coil it is best to apply the following formula:

$$L_{(\text{microhenries})} = \frac{0.01 D w^2}{\frac{l}{D} + 0.44},$$

where D and l are, respectively, the diameter and length of winding in centimetres (Fig. 31*a*), and w stands for the number of turns.

Example 3. Find the inductance of a coil in which $D = 2$ cm, $l = 4$ cm, and $w = 50$ turns.

Solution.

$$L = \frac{0.01 \times 2 \times 50^2}{\frac{4}{2} + 0.44} = \frac{50}{2.44} = 20.5 \text{ microhenries.}$$

If the value of inductance is known, the following formula is applied to find the number of turns, having first assumed definite values of l and D:

$$w = 10 \sqrt{\frac{L\left(\frac{l}{D} + 0.44\right)}{D}}.$$

Having determined the number of turns, we next find the diameter of insulated wire:

$$d_{ins} = \frac{l}{w}.$$

It is desirable to proportion coil dimensions in such a way that the relation $\frac{l}{D}$ has a value between 0.2 and 2 (the optimum value, at which Q of the coil is maximum, is 0.4).

Example 4. Determine the number of turns and diameter of wire for a 10-microhenry coil, in which $l = 3$ cm and $D = 2$ cm.

Solution.

$$w = 10\sqrt{\frac{10\,(1.5 + 0.44)}{2}} = 33 \text{ turns};$$

$$d_{ins} = \frac{30}{33} = 0.9 \text{ mm}.$$

If 0.9 mm wire is not available and we are forced to use a thinner wire, the latter must be wound spaced, so that the winding length is 3 cm, as required by the conditions of the problem.

The inductance of multi-layer coils is found with the help of the following formula:

$$L_{\text{(microhenries)}} = \frac{0.08D^2w^2}{3D + 9b + 10c},$$

where D, b and c denote the average diameter, length and depth of winding in centimetres respectively, while w signifies the number of turns (see Fig. 31*b*). If the maximum and minimum diameters of the coil (D_{max} and D_{min}) have been measured, the average diameter D is found as follows:

$$D = \frac{D_{max} + D_{min}}{2}$$

also:

$$c = \frac{D_{max} - D_{min}}{2}.$$

The formula given above cannot be used to find the number of turns for a stipulated inductance value, as in this case the values of D and c are unknown. The number of turns can be determined with the help of the following approximate formula for the inductance of a multi-layer coil:

$$L_{\text{(microhenries)}} = 0.01\, D_{min}w^2.$$

This formula gives a somewhat lowered value of L. Transposing the formula, we have:

$$w = 10\sqrt{\frac{L}{D_{min}}}.$$

The number of turns given by this formula will, in reality, be too large, and after the coil has been wound it will be necessary to trim the inductance by removing several turns from the coil.

Example 5. Find the inductance of a 200-turn multi-layer coil of the following dimensions: $D_{min} = 1.4$ cm, $D_{max} = 2.6$ cm, $b = 1.1$ cm.

Solution.

$$D = \frac{1.4 + 2.6}{2} = 2 \text{ cm}; \qquad c = \frac{2.6 - 1.4}{2} = 0.6 \text{ cm}.$$

then:

$$L = \frac{0.08 \times 2^2 \times 200^2}{3\times 2 + 9\times 1.1 + 10\times 0.6} = \frac{12{,}800}{22} = 580 \text{ microhenries}.$$

The approximate formula gives:

$$L = 0.01 \times 1.4 \times 200^2 = 560 \text{ microhenries}.$$

Example 6. Find the number of turns of a multi-layer coil which must have $L = 200$ microhenries when $D_{min} = 2$ cm.

Solution.

$$w = 10\sqrt{\frac{200}{2}} = 100 \text{ turns.}$$

It should be taken into consideration that placing a shield over a coil reduces *the value of L by 10-20%. Hence, when computing the number of turns for a shielded coil, allow additional 10-20% for the value of L.*

16. QUESTIONS AND PROBLEMS

1. What are free oscillations and what are their properties?
2. Explain the nature of oscillatory process in a tuned circuit.
3. What do the amplitude and frequency of free oscillations in a tuned circuit depend upon?
4. Why are free oscillations always damped?
5. Why cannot oscillations take place in a circuit consisting of a capacitor and ohmic resistance?
6. What is the quality factor of a tuned circuit and what is its value for well-designed circuits?
7. How is the 90° phase shift between current and voltage in an oscillatory circuit explained?
8. What happens to the energy in an oscillatory circuit $\frac{1}{8}$-th of a period after the capacitor has begun to discharge through the coil?
9. What will happen to the frequency and wavelength of free oscillations in a tuned circuit if the inductance is decreased nine times and at the same time the capacitance is reduced four times?
10. Draw a tuned circuit employing a fixed capacitor, a continuous-tuning variometer and an additional coil supplied with taps and a switch for coarse tuning.
11. What has to be done to make the free oscillations in a tuned circuit continuous?
12. A tuned circuit has a natural frequency f_0 of 2,000 kc. This circuit is connected to an alternator (alternating current generator) with a frequency f equal to 2,500 kc. What will be the frequency of oscillations in the tuned circuit?
13. What is the difference between free and forced oscillations?
14. Explain the resonance phenomenon.
15. What has to be done to obtain maximum possible values of voltage across coil and capacitor on series resonance?
16. Is it possible to obtain series resonance in a tuned circuit without connecting an alternator into the circuit?
17. What does the resonance curve indicate?
18. A series resonance takes place in a tuned circuit. In this circuit $x_L = x_C = 800$ ohms, $r = 40$ ohms. The alternator voltage U is 10 volts. Determine the value of current flowing through the tuned circuit and find the value of voltages appearing across L and C.
19. What is the impedance offered by a tuned circuit to the alternator in the case of parallel resonance, if the tuned circuit has the following parameters: $L = 150$ microhenries, $C = 200$ picofarads, and $r = 25$ ohms?
20. Define coupled circuits.
21. Why do the coils of high-frequency circuits use cores made of magnetic dielectric materials and not of steel?
22. What linking of two coupled circuits do we call critical coupling?

23. Describe the peculiarities of resonance curves of coupled circuits.
24. Why do the shields placed over inductance coils increase the ohmic resistance of tuned circuits?
25. Is it a good practice to place a solid copper shield between the coils of a high-frequency transformer for eliminating parasitic capacitive coupling?
26. What will happen to the frequency of natural oscillations in a tuned circuit if the number of turns of inductance coil in it is increased?
27. What measures should be taken to increase the sharpness of resonance in a tuned circuit?
28. By how much should the inductance of a tuned circuit be changed to increase its wavelength 4 times?
29. Why will the inductive coupling between two coils be at its lowest value when the coil axes are at right angles to each other?
30. How is the bandwidth of a tuned circuit determined from the resonance curve of the circuit?
31. Why does the coil inductance decrease when a shield is placed over the coil?
32. What is the effect of a core made of magnetic dielectric material upon the inductance of a coil?
33. Why are trimming capacitors used in tuned circuits?
34. How should an inductive coupling be arranged between tuned circuits placed at a considerable distance from each other? Draw the circuit of the whole arrangement.
35. What should be the bandwidth of a tuned circuit for various types of radio transmissions?
36. One tuned circuit is adjusted to resonate at 300 metres, and another tuned circuit at 6 metres. The quality factor is 50 in both cases. Find the bandwidth of each of the two tuned circuits.
37. What is the effect of a shunting resistor, connected across a tuned circuit, upon the quality and the resonant properties of the circuit?
38. How does the internal resistance of a generator, feeding a parallel-resonant tuned circuit, affect the resonant properties of the circuit?
39. Define reflected impedance.

CHAPTER III

AERIALS AND PROPAGATION OF RADIO WAVES

17. ELECTROMAGNETIC WAVES

When a high-frequency alternating current flows in a conductor, electric and magnetic fields are set up around it. These fields form *an electromagnetic field, which propagates in all directions from the conductor with the speed of 300,000 kilometres per second.*

An electromagnetic field moving through space is otherwise called an electromagnetic wave. Radio wave is the other name for such a wave.

Electric and magnetic fields cannot exist independently of each other. *Any change of an electric field gives rise to a magnetic field. Conversely, any change of a magnetic field creates an electric field.*

It is not correct to attach the name of "electromagnetic fields" to *constant* electric and magnetic fields existing in some parts of space. Such fields act independently of each other, and there is no interaction between them. An actual electromagnetic field is always a combination of two equally important alternating fields, one of which is magnetic and the other electric. These two fields interact and support each other.

Such interaction between magnetic and electric fields accounts for the propagation of the electromagnetic field in the space. A changing electric field creates a correspondingly changing magnetic field. But the changing magnetic field creates, in its turn, a new electric field which also changes, produces a new magnetic field, etc.

Thus an electromagnetic field represents an oscillatory process extending to new parts of space, one after another.

During its propagation an electromagnetic field parts from the conductor around which it was first created. The current feeding this conductor may now be switched off, but the electromagnetic wave will continue on its way through the space.

Electromagnetic waves carry the energy received by them from the current flowing in the conductor. Thus a wire carrying alternating current radiates radio waves. These waves are propagated in all

directions just like light waves, which, incidentally, are also a type of electromagnetic waves.

Hence it is customary to say that *wires carrying alternating currents radiate electromagnetic waves through space*. The higher the *power of the alternating current in a wire, the greater is the energy* of the radiated waves. This energy also depends to an even greater extent upon the frequency. If the frequency of the radiated wave is increased 2, 3, 4, etc., times, the energy contained in the wave increases 4, 9, 16, etc., times.

Low-frequency currents produce only an insignificant amount of radiation in comparison with high-frequency currents of higher frequencies. Sufficiently strong radiation, and hence successful radio transmission over long distances, are obtained through the application of currents with frequencies between hundreds of thousands and many million cycles per second.

Electromagnetic waves need no atmosphere for their propagation and travel easily through interplanetary space. However, it would be wrong to say that electromagnetic waves are nothing but transfer of energy through the void. Nature knows no void, and energy can not exist without matter. In the light of the latest scientific attainments we can now assert that *electromagnetic waves represent moving matter*.

In the past some scientists assumed that the space which is free of air, as well as the gaps between the particles of all the substances, was filled with a special kind of matter—"space ether", and this led many to regard radio waves as waves in the "ether".

Modern physics has repudiated the existence of the "ether", but many people have fallen into the habit of saying that "radio stations radiate ether waves", that "the radio waves propagate through the ether", etc.—and this habit is, apparently, hard to break.

18. THE AERIAL AS AN OPEN OSCILLATORY CIRCUIT

An aerial is a system of wires used for the transmission of radio waves by transmitting stations and for the reception of waves by receiving installations. In other words, an aerial converts high-frequency alternating current energy into the energy of radio waves, and, conversely, converts the radio wave energy into high-frequency alternating current energy.

A. S. Popov was the first man in the world to employ an aerial. As the years went by further contributions to the theory and practice of aerials were made by M. V. Shuleikin, A. A. Pistolkors, M. A. Bonch-Bruyevich, A. L. Mints, G. Z. Aisenberg and other Soviet scientists and engineers. Many original types of aerials were thus created in the Soviet Union.

Let us examine the design of various aerials and their operating principle.

A closed oscillatory circuit of limited dimensions in comparison *with its wavelength radiates electromagnetic waves very poorly. This can be explained as follows.*

Electromagnetic waves are radiated by a conductor carrying high-frequency current. If such a conductor is bent to form a loop, as shown in Fig. 32*a*, the currents through its two halves will flow in opposite directions.

Waves generated by the two halves of the conductor will also be of opposite phases and, if distance *d* between the two halves is too small in comparison with the wavelength, these waves will cancel each other out in space.

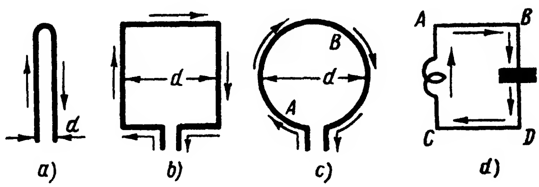


Fig. 32. Opposite directions of currents through various parts of an oscillatory circuit

Thus a conductor shaped as a loop radiates practically no electromagnetic waves. The same can be said about a conductor shaped as a rectangular or circular wire turn (Fig. 32*b* and *c*) whose dimensions are much smaller than the wavelength. Currents in the opposite sides of rectangular turn are flowing in different directions and therefore the waves set up by these currents are of opposite phases. These waves completely cancel each other out in the direction perpendicular to the plane of the turn. The phase shift between these waves in the direction along the turn plane is somewhat different from 180°, since one of the waves travels a certain extra distance equal to one side of turn *d* and, because of this, experiences a certain phase lag. However, if the side of the turn is much smaller than the wavelength, such a lag is quite negligible and it can be practically considered that the waves travelling in these directions are also mutually cancelled.

When the diameter of a circular wire turn is small, each element of the conductor, for instance element *A* in Fig. 32*c*, has a correspondingly diametrically opposite element *B*. In such elements the currents flow in opposite directions. Obviously the waves generated by these elements have opposite phases and practically cancel each other.

If the dimensions of turn *d* were equal to a considerable portion of wavelength λ, the waves travelling along the turn plane from its opposite sides would have a phase shift quite different from 180°,

because one of the waves would be considerably lagging and no mutual cancelling of waves would take place. Only in the direction perpendicular to the turn would the waves still be travelling similar *distances and cancelling each other.*

In oscillatory circuits used by radio equipment operating on medium and short waves the coil turns usually have a diameter of several centimetres, while the wavelength is measured in tens and hundreds of metres. When such a relation exists, it may be assumed that each turn, taken individually, does not radiate and, hence, the whole coil is also not radiating.

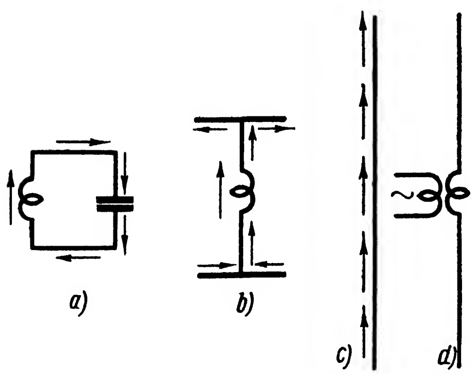


Fig. 33. Conversion of a closed oscillatory circuit into an open one

On these waves the whole tuned circuit can be considered as a single turn, in the opposite elements of which the currents are flowing in different directions. As may be seen in Fig. 32*b*, the currents in conductors *AB* and *CD* are flowing in opposite directions. The same can be said about the currents in sections *AC* and *BD*, i. e., in the coil and the capacitor. Since the geometrical sizes of a tuned circuit are usually small in comparison with the wavelength, it may be assumed that such a circuit gives off a very weak radiation.

It is, however, possible to change the design of an oscillatory circuit in such a way that the currents in its separate elements will have a similar direction in space, i.e., the oscillations in separate elements of the tuned circuit will coincide in phase. Then the waves generated by such elements will not cancel each other, and a considerable radiation will take place. This is obtained by the conversion of a closed oscillatory circuit (Fig. 33*a*) into *an open oscillatory circuit—the aerial.*

If the plates of a capacitor are pulled far away from each other and the wires connecting the coil with the capacitor are stretched out to form a straight line (Fig. 33*b*), the current directions in these wires will be the same. Such a tuned circuit, however, does not sufficiently radiate since no radiation is yet given off by the coil, and the currents flowing along the capacitor plates are still opposite to each other in direction and are at right angles to the currents in connecting wires.

A further considerable increase in the intensity of radiation may be obtained by stretching the coil wire into a straight line and using, instead of the plates of capacitor, straight wires of sufficient length

(Fig. 33*c*). In this case the current directions in all the elements of the wire will be the same, i. e., the oscillations in all parts of the *wire will be of the same phase and the radiation intensity maximum. It follows that an open tuned circuit, in its simplest variant, is* represented by a straight wire. In practical cases a small coil L_{coup} is still included in such a wire for coupling with the generator (Fig. 33*d*).

Every conductor possesses a certain amount of natural inductance and capacitance distributed along its length and, accordingly, may

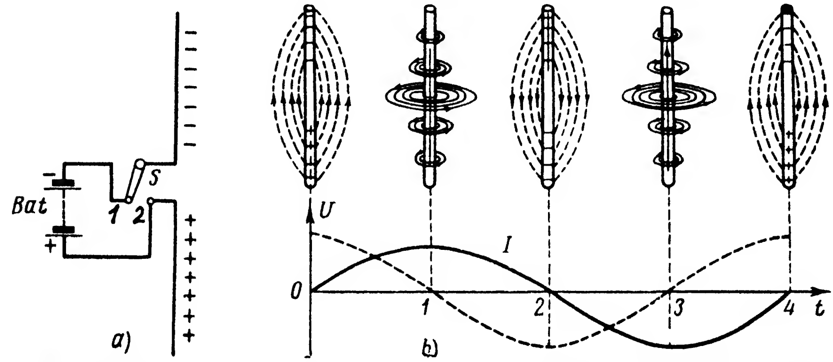


Fig. 34. Free oscillations in an open circuit, and the oscillatory process

be considered an oscillatory circuit with its own definite frequency. As in any oscillatory circuit, free electrical oscillations may be set up in a wire. In Fig. 34*a*, when switch *S* is set to position *1*, both halves of the wire are charged with opposite polarity from battery *B*. If the switch is shifted to position *2*, the electrons will move along the wire from the lower half to the upper one and then in the opposite direction, thus setting free damped oscillations in the wire. As in a closed tuned circuit, such oscillations will exist because of the inductance and capacitance of the wire.

Separate phases of the oscillatory process in the wire are shown in Fig. 34*b*. The upper part of the drawing shows the distribution of electric and magnetic fields, while the lower part gives the curves of current and voltage changes in the aerial.

The voltage at any point of the aerial is the potential difference between the given point and the point symmetrically opposite to it on the other half of the wire.

It should be kept in mind that the current curve also indicates the change of magnetic field intensity with time, and the voltage curve—the electric field intensity change. In Fig. 34*b* the voltage curve and corresponding electric field are indicated by a broken line, while the current curve and the respective magnetic field—by a solid line.

At the initial moment (point *0* in Fig. 34*b*) the wire carries the potential energy of the electric field of charges concentrated in the upper and lower halves of the wire. There is no current as yet, but the potential difference is at its maximum value. As the electron movement commences along the wire, the current begins to increase and the voltage to decrease. The electric field energy is gradually converted into the kinetic energy of the magnetic field set up by the current. One-quarter of a period later the electric field is replaced by the magnetic field. At this moment (point *1* in Fig. 34*b*) the current is at its maximum value while the voltage is equal to zero, following which the current and the magnetic field begin to decrease. This gives rise to e.m.f.

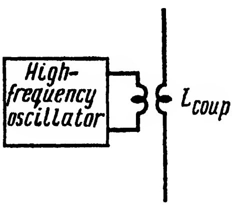


Fig. 35. Inductive coupling of an open tuned circuit with an oscillator

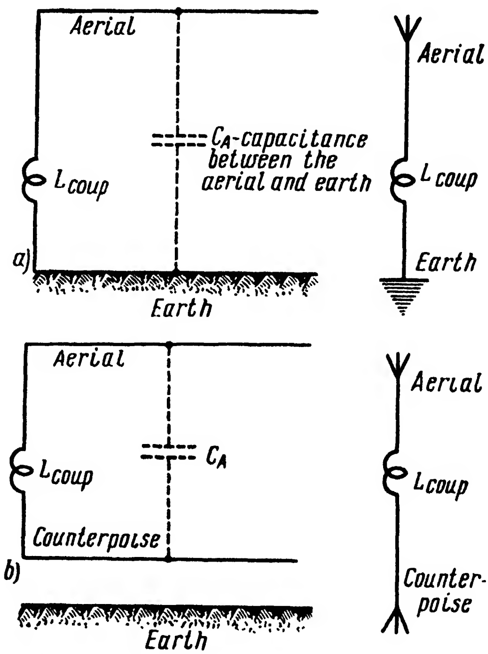


Fig. 36. An aerial employing an earth connection *(a)* or a counterpoise *(b)*

of self-induction which keeps the electrons moving, thereby recharging the wire. The energy is transferred from the magnetic field into the electric field. At the end of the second quarter of the period (moment *2*) the energy is again concentrated in the electric field, but the direction of the field is reversed. During the following half of the period the whole process is repeated in reverse order, and the initial state is restored. At intermediate moments, not shown in the upper drawing, the electric and magnetic fields will exist simultaneously, since the oscillatory energy is distributed between the two fields. As may be seen in Fig. 34*b*, both the electric and magnetic fields are present along the entire conductor, and the magnetic field will be most intense in the middle of the wire where the current is at its maximum value. The intensity

of the magnetic field will be zero at the ends of the wire because of the absence of current at the wire ends.

An open oscillatory circuit, represented by a straight wire in which free electrical oscillations take place, is referred to as a *symmetrical aerial* or simply a *dipole*. To obtain continuous oscillations in a dipole it must be coupled to a generator (oscillator). Such a coupling may be, for instance, inductive (Fig. 35).

In practical cases the aerial of a transmitting radio station operating on long, medium and sometimes on short waves is made as follows. The aerial itself, i.e., a wire system playing the role of one of the capacitor plates, is suspended over the ground as high as possible. The second plate of the capacitor is represented by the ground or else by another wire, called a counterpoise, suspended at a short distance above the ground and insulated from it. Such designs are typical of non-symmetrical aerials. Capacitance C_A between the aerial and ground (or counterpoise) reaches several tens or even hundreds of picofarads. Fig. 36*a* and *b* shows the schematic of an aerial employing an earth connection and a counterpoise.

The same figure shows the symbols for aerial, earth and counterpoise, as used in radio circuits.

19. VOLTAGE AND CURRENT DISTRIBUTION IN AN AERIAL

Unequal distribution of current along the wire is an important peculiarity of an open oscillatory circuit. It is in this respect that open oscillatory circuits differ from tuned circuits of the closed type. The current is zero at the wire ends in an open circuit because there is no place for the electrons to go to or to come from at the ends of the conductor. The current increases towards the centre of the wire and reaches a maximum value in the wire middle, where the coil coupling the aerial to the generator is located. It is customary practice to represent the distribution of current in an aerial graphically, as shown by the solid curve *1* in Fig. 37. In such a representation the current at any point of the aerial is determined by the distance from the point to curve *1*.

The points at which the current is zero (in the given case the ends of the aerial) *are called current nodes, while the points where the current is maximum* (centre of the wire) *are called current loops.*

The voltage in the aerial is also distributed unequally.

Fig. 37 shows (broken line) the distribution of voltage in the aerial; here the positive polarity of voltage is represented by the curve at one side of the wire, and the negative polarity at the other side. The maximum value of voltage, i.e., *the voltage loop*, is always located at the wire ends. In the middle point, where the coupling coil is located, the voltage value is zero (voltage node). Voltage loops

occur at current nodes and, conversely, a voltage node always corresponds to the point where a current loop takes place. One-half of a wavelength is placed along the wire, while the curves representing the distribution of current and voltage are shifted in phase one-quarter of a wavelength in respect to each other.

Let us analyse in greater detail the oscillatory process taking place in an aerial. The maximum value of voltage corresponds

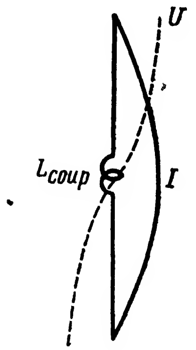


Fig. 37. Distribution of current and voltage in an open oscillatory circuit (in a dipole)

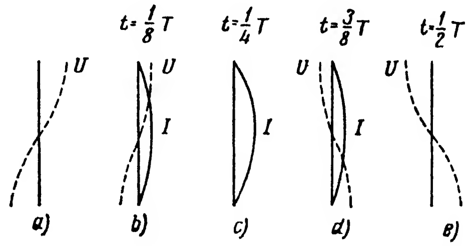


Fig. 38. Distribution of current and voltage in an open oscillatory circuit at different periods of time

to the maximum potential energy of electrical charges. These charges are the greatest at the wire ends. When the electrons begin to move, the energy of the charges decreases, the current increases and the kinetic energy of the current becomes larger. The maximum value of this energy (current maximum) will not occur at the place where the charges are concentrated but rather at the point on the wire where the maximum quantity of electrons are moving, i.e., in the middle.

Therefore the maximum values of current and voltage do not coincide in either position or time. There is a quarter-period $\left(\frac{T}{4}\right)$ phase shift between voltage and current in an aerial, just as in a closed oscillatory circuit. Fig. 37 shows the current and voltage distribution at a certain moment, while Fig. 38 gives the curves of current and voltage distribution in a conductor at different moments during one-half of a period.

When an oscillation begins (Fig. 38*a*), there is no current in the aerial while the voltage is at its maximum value. One-eighth of a period later the voltage will have decreased and the current will have appeared in the wire (Fig. 38*b*). One-quarter of a period after the oscillation has begun the current reaches its maximum value, while the voltage becomes zero (Fig. 39*c*). The current then decreases and the voltage reappears. However, the voltage now has opposite polarity, bacause the two halves of the wire are charged differently (Fig. 39*d*). One-half of a period after the oscillation has begun the

current decreases to zero and the voltage reaches its maximum value (Fig. 39*e*). The process is then repeated in the opposite direction.

The phenomena observed in the aerials may be explained on the basis of electromagnetic wave distribution along the wire. If a high-frequency alternator is connected to a single-wire or a double-wire line, a so-called *travelling wave* of current and voltage will move along the line with the speed of 300,000 km per sec. This wave is an electromagnetic wave, because magnetic and electric fields will form around

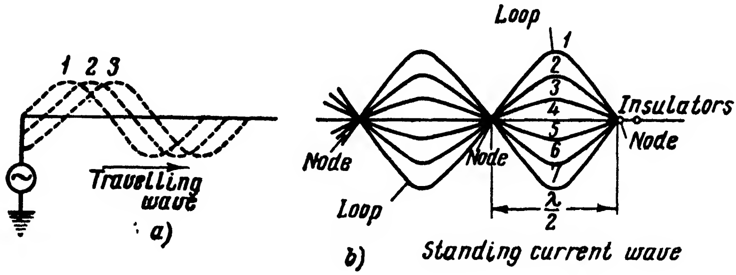


Fig. 39. Graphic representation of travelling and standing waves in a wire for several different moments of time

any wire carrying current and voltage. When a travelling wave is present, the current and voltage at any point of the wire will be constantly changing. Therefore it is possible to show graphically voltage and current distribution along the wire only for some definite moment. Fig. 39*a* shows the distribution of current and voltage of a travelling wave for three different moments in sequence.

The creation of a travelling wave may be demonstrated by an experiment with long rope. If one end is fixed to a stationary object and the other end is given a shake, a travelling wave will move along the rope.

When a travelling wave with voltage and current reaches an obstacle, e.g., a dielectric insulator at the end of the wire, it is reflected and begins to move in the opposite direction to the direct wave. These two waves meet and produce a special distribution of current and voltage, known as a *standing wave.* A standing wave can be also obtained in the experiment with the rope by continuously swinging the free end of the rope and sending travelling waves to the attached end, these waves reflecting to meet the following travelling waves.

When a standing wave is present in the aerial the amplitudes of current and voltage are different at various points on the wire, but the distribution of current and voltage amplitudes along the wire length remains stationary and does not change with time. Only the instantaneous values of current and voltage change in such a case. Hence the distribution of current in a standing wave may be represented for different moments, as shown in Fig. 39*b*. For the case of a standing wave the presence of voltage and current loops

and nodes is typical. The mutual location of such loops and nodes was shown in Fig. 37, and the change of current and voltage with time for a dipole was illustrated by the curves of Fig. 38.

Unlike a closed oscillatory circuit, an open circuit is noted for possessing both a natural frequency and harmonics of such frequency. This means that if the aerial is excited by an oscillator and the frequency of the oscillator is varied, the condition of resonance will be observed not only on the fundamental frequency but also on higher harmonics whose frequencies are multiples of it.

This property is typical for all oscillatory circuits with distributed parameters, such oscillatory systems differing in the given respect from those with lumped parameters. Taking, for instance, a tightly stretched string with a certain mass and elasticity distributed along its entire length, it can be shown that it is possible to make such a string oscillate on a fundamental frequency and on harmonics—which cannot be done with a pendulum.

20. NATURAL FREQUENCY AND THE WAVELENGTH OF AN AERIAL

The frequency of free oscillations of an open tuned circuit depends upon its capacitance and inductance. If the full length of the aerial wire is known, these parameters can be approximately determined. Every metre of wire has a capacitance of about 5 pf and an inductance of about 2 μh. For instance, if the full length of an aerial is 40 metres, then the aerial capacity $C_A = 5 \times 40 = 200$ pf and its inductance $L_A = 2 \times 40 = 80$ microhenries. This calculation holds true only for an earthed aerial consisting of a single wire.

The longer the aerial wire, the larger are its inductance and capacitance and, consequently, the lower is the frequency (and the longer the wavelength) corresponding to the free natural oscillations of the given aerial. Considering that the speed of current distribution along the wire is 300,000 km per sec, equations for obtaining the natural wavelength or frequency of an aerial may be derived.

In one-half of a period electric current flows along the wire in one direction only. Hence, the wire length l of an open tuned circuit is given by $\frac{\lambda}{2}$, which is already known from the study of current and voltage distribution in an aerial wire. This gives the following expression:

$$\lambda = 2l.$$

Therefore the wavelength of natural oscillations in an aerial insulated from the earth is equal to twice the length of the wire.

Substituting frequency for wavelength in the equation, by using the following expression

$$\lambda_{(\text{metres})} = \frac{300{,}000}{f_{(\text{kc})}},$$

we have:

$$\frac{300{,}000}{f_{(\mathrm{kc})}} = 2l_{(\mathrm{metres})}$$

or:

$$f_{(\mathrm{kc})} = \frac{150{,}000}{l_{(\mathrm{metres})}}.$$

For instance, if the length of the aerial wire is $l=30$ m, then the natural frequency is given by: $f = \frac{150{,}000}{30} = 5{,}000$ kc, while the wavelength is equal to $\lambda = 2\times 30 = 60$ m.

It should be noted that owing to the influence of the earth and other nearby objects the actual natural wavelength is, in practice, somewhat longer than the doubled length of the wire.

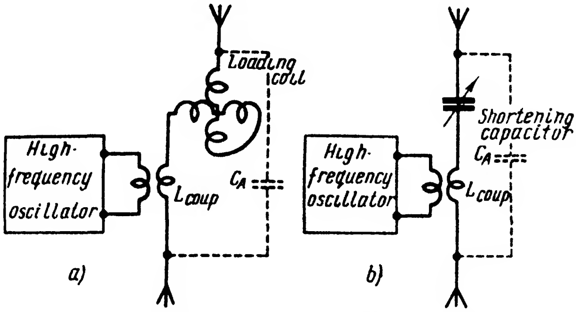


Fig. 40. Lengthening and shortening of the aerial wavelength

Maximum intensity of radiation by an aerial requires strong oscillations in the aerial wire. If the aerial is fed with high-frequency current from an alternator, forced oscillations take place in the wire. The amplitude of such oscillations is greatest at resonance, when the alternator frequency is equal to the natural frequency of the aerial. In the case of inductive coupling to the alternator, the aerial serves as a secondary tuned circuit, in which, as we already know, only series resonance can take place.

Since a radio transmitter is usually adjustable to several wavelengths, the aerial must be capable of tuning to respective frequencies. The latter can be secured by varying the length of the aerial wire to change its natural frequency. Such a method is impractical in most installations, and aerials are usually tuned by means of variable capacitors or variometers.

Lengthening of the natural wavelength of an aerial requires the series connection of a coil into the aerial, which, in effect, is equivalent to lengthening the aerial wire. A variometer can serve as the coil and ensure continuous adjustment over a required frequency range (Fig. 40*a*). *Series connection of a capacitor into the aerial shortens the natural wavelength of the aerial wire* (Fig. 40*b*). This is quite obvious,

because the capacitor, in the given case, is connected in series with capacitance C_A of the aerial itself, thus reducing the total capacitance of the tuned aerial circuit. When a series-connected coil or a capacitor, taken separately, fail to give a required frequency coverage, they are sometimes both connected into the aerial.

The capacitor and coil mentioned above are called a shortening capacitor and a loading coil.

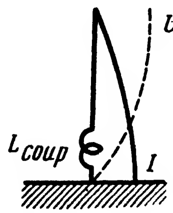


Fig. 41. Distribution of voltage and current in an earthed aerial

It should be noted that in an earthed-type aerial there can be no voltage loop at the end of the earthing wire, because the earth connection is always at zero potential. Hence at the point where the earthing lead is connected to earth there is always a voltage node and a current loop. This is why the length of an earthed aerial is equal to one-quarter of the wavelength when the aerial is adjusted to operate on the fundamental wave (1st harmonic). The distribution of current and voltage in this case is shown in Fig. 41. It follows that we can write a mathematical relation giving the dependence of natural wavelength of an earthed aerial upon the length of the aerial wire:

$$l = \frac{\lambda}{4} \text{ or } \lambda = 4\,l \text{ (for fundamental wavelength).}$$

Because of the influence of the earth and nearby objects, the actual natural wavelength is usually longer than the above equation would indicate. In practice the coefficient by which l must be multiplied to give λ is usually taken as from 5 to 6.

21. RECEIVING AERIALS

In general all the aerials may be classified into *receiving* and *transmitting* aerials. In most cases there is no fundamental difference between the two classes of aerials; receiving aerials can be used for transmission purpose and vice versa.

On reception, radio waves act upon the aerial wire (or upon any conductor, for that matter) and generate electromotive force in it. The electromagnetic field of the radio wave conveys oscillatory motion to the electrons of the given wire, causing electrical current to flow in it. The frequency of the current thus generated in the receiving aerial will be the same as the frequency of current in the aerial of the transmitting station being received.

Receiving aerials are usually not tuned to the frequency of signals being received; therefore, the length of aerial wire is of no considerable importance in such cases. The voltage set up in receiving aerials

by incoming waves is negligibly small, and the problem of insulating such aerials is relatively simple.

In any aerial, in any electrical circuit, energy losses take place. Such losses are of no great significance in receiving aerials designed for modern high-gain radio receivers. These receivers secure good reception even with poor aerials. Only in case of simple low-gain receivers it is desirable to use good aerials with small energy losses.

The best reception is obtained with external (outdoor) aerials. In the open country the average height of an aerial above the ground should be 6-10 metres; in cities with tall buildings it should be 3-5 metres above the roof. Receiving aerials are usually made of bare copper wire or a special stranded wire, with a diameter of 1-2 mm in either case. In lieu of copper wire, steel wire may be used in case of urgency. Insulated wires may also be used as aerials since insulation does not present an obstacle to radio waves.

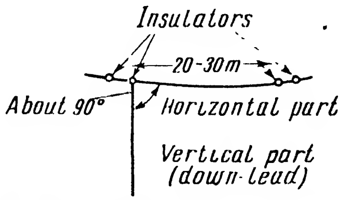


Fig. 42. Single-wire L-type aerial

The following types of aerials are used for the reception of broadcasting stations.

Single-wire L-type aerial. This is comprised of a single wire with a shape resembling an inverted letter L (Fig. 42). The ends of the horizontal part of the aerial must be suspended on insulators to prevent leakage of signal currents to earth through the guys and aerial masts. Signals picked up by the aerial reach the radio receiver over the lead-in wire, which should also be well insulated.

Single-wire T-type aerial. This type is rigged up when the radio receiver is conveniently located between two high points of suspension. The lead-in wire is connected to the middle point of the horizontal part of the aerial.

Vertical or tilted aerial. This has no horizontal part and consists of a single vertical or tilted wire. In some cities a single central radio mast is sometimes installed on a tall building, the top of the mast supporting the ends of tilted aerials. More than one radio receiver should not be connected to one aerial, as they would interfere with each other.

Lumped-capacitance aerial. This differs from the vertical aerial in that its top end is provided with a conductor shaped as a spiral or a wire brush. The conductor is intended to increase the capacitance, but such an aerial has no particular advantages over other types.

Indoor and artificial aerials. These are employed in cases when rigging up an external aerial is difficult or impossible. Indoor and artificial aerials give poorer reception than external aerials, and when used in a city they pick up more interference than the outdoor

(external) type. The higher the room in which an indoor aerial is installed, the better is the reception. The walls of reinforced concrete buildings readily absorb the radio waves, and the installation of indoor aerials in such buildings is hardly worthwhile.

The role of artificial aerial may be played by some conductor which is simultaneously used for some other purpose having no relation to radio reception, e.g., power wiring. When a power line is used as an artificial aerial, one of its wires is connected to the radio receiver through a 50-100 pf isolating capacitor (the exact value of capacitance is of no importance). To assure safety, a fuse rated not in excess of 0.25 ampere should be connected in series with the connecting lead (Fig. 43). Power line current, because of its low frequency, is practically blocked by the small capacitor. However, high-frequency currents set up by radio waves in the power wiring easily reach the radio receiver through the capacitor.

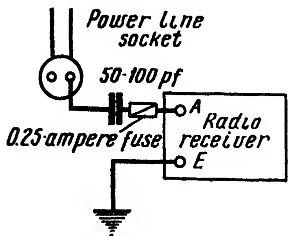


Fig. 43. Power line used as an artificial aerial

It is not a good practice to use power wiring as an artificial aerial, because careless handling of the equipment can result in damage to the radio receiver and to the wiring. Besides, all types of interference reach the receiver via the power wiring, making good reception difficult.

Modern multi-valve radio receivers are extremely sensitive and provide good reception of distant stations even when using a short piece of wire instead of a full aerial. But, even these receivers should be used with a good external aerial wherever possible, because they give a decidedly improved performance with such an aerial.

The field of an electromagnetic wave acting upon a receiving aerial is evaluated in terms of its *intensity*. This field intensity is taken as the potential difference in 1-metre length of electrical lines of force. Distant radio stations give field intensities of only several microvolts per metre (μv/m) in receiving localities. If the field intensity of a radio station in a given locality is 10 μv/m, this is equivalent to the field intensity in a capacitor whose plates are spaced at one metre from each other and are charged up to 10 microvolts. Stronger field intensities of radio waves are measured in millivolts per metre (mv/m).

The electromotive force set up in an aerial by radio waves depends upon the field intensity, the height of the receiving aerial and its design. In order to compare the relative quality of aerials having different height and shape, we speak of *the effective height of an aerial*, which is usually smaller than its geometric height.

If the effective height of an aerial is denoted by h_e and the field intensity by E, then the electromotive force E_A set up by a radio

wave in an aerial can be found by the following simple equation:

$$E_A = Eh_e.$$

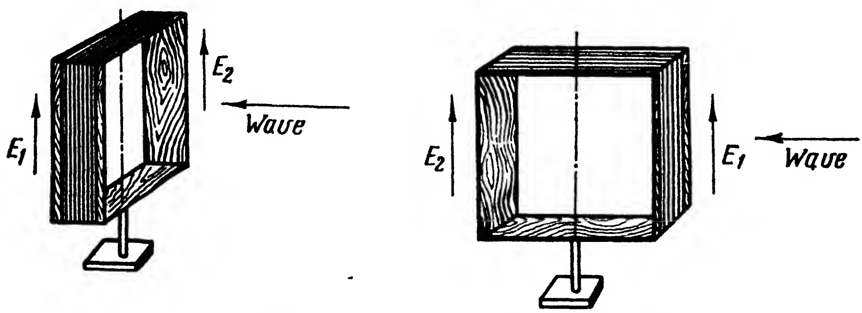
For example, if $h_e = 8$ m and $E = 50$ μv/m, the electromotive force E_A equals 400 μv.

A vertical aerial operating on its natural wavelength has an effective height equal to approximately 0.64 of its geometric height h. If such an aerial employs a loading coil of considerable inductance, the effective height of the aerial will be decreased to about 0.5 h. If the aerial is supplied with a horizontal part, the value of h_e can reach 0.8 h and even 0.9 h.

22. LOOP AND MAGNETIC AERIALS

A loop aerial has some interesting features. It possesses directional properties and, constructionally, represents a large-dimension inductance coil. It is most sensitive to signals arriving along its plane; the loop does not pick up signals which reach it perpendicular to its plane. Turning the loop aerial secures the best audibility of a desired radio transmitting station and helps to eliminate interference from radio stations located in other directions.

The directional characteristics of a loop aerial are attributed to the following. If the loop plane is set at right angles to the direction of incoming radio waves, as shown in Fig. 44a, two equal and opposite



$E_1 = E_2$; $E_{common} = E_1 - E_2 = 0$
a)

$E_1 \neq E_2$; $E_{common} = E_1 - E_2 \neq 0$
b)

Fig. 44. Directional characteristics of a loop aerial

electromotive forces E_1 and E_2 will be set up in both halves of each wire turn of the loop. As a result, the total e.m.f. in the whole of the loop will be zero. However, if the loop is so set that its plane is parallel to the direction of the incoming wave (Fig. 44b), E_1 will no longer be equal to E_2 because their phases are not the same, for the wave reaches one half of the turns before the second half.

A resultant e.m.f. will then appear in the loop aerial, the value of such an e.m.f. being directly proportional to the size of the loop aerial and the number of turns in the loop, and inversely proportional to the length of the wave.

The directional property of a loop aerial is characterised by a very sharp reception minimum (sudden disappearance of the signal) and a rather broad maximum.

In recent times the *magnetic aerial* has come to the fore. This is a version of the loop aerial. Constructionally, a magnetic aerial is represented by a coil with a ferromagnetic core, usually made of ferrite. Owing to the high magnetic permeability of ferrite (several hundred units), the magnetic field of a radio wave sets up in the coil a magnetic flux of higher intensity than that obtained without the core. Hence even a small coil installed in a magnetic aerial gives rise to an e.m.f. of a value similar to that obtained in loop aerial of much larger dimensions.

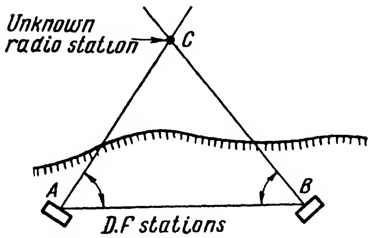


Fig. 45. Determining the location of an unknown radio station

In practical constructions the core of a magnetic aerial has a diameter of 15-30 cm, and its cross-sectional area is equal to 1-2 sq.cm. The directional effect in a magnetic aerial is the same as that in a loop aerial. A magnetic aerial is usually placed inside the radio receiver. To make use of its directivity, the whole receiver has to be rotated or a special aerial-turning handle has to be provided. The first approach is used in portable receivers, and the second in larger sets designed for stationary operation.

Loop or magnetic aerials are generally used when it is required to reduce interference from other radio stations operating on wavelengths close to that of the desired station but located in other directions.

Radio direction finding is an important application of the loop aerial. Here special radio receivers equipped with loop aerials are used to determine the location of transmitting radio stations (Fig. 45).

A direction finding system works as follows. During reception of signals of a radio station whose location must be determined, loop aerials of d.f. stations A and B are rotated until the signal disappears (remember that a loop aerial has a very sharp minimum of reception). The angles between straight line AB and the directions at the unknown station are then measured. The crossing of lines AC and BC on the map at point C gives the location of the transmitting station. Ships and aircraft employing direction finding determine their position with sufficient accuracy in foggy weather, in clouds and at night.

Another navigational aid employs large-sized loop aerials for transmission. Such installations, used as *radio beacons, radiate* directional signals, helping ships and airplanes to follow required courses. Directional transmissions with the loop aerials have also found other applications.

23. EARTHING FACILITIES AND COUNTERPOISE

In radio equipment earthing arrangements serve a double purpose. The earth, together with the aerial, conveys radio signals to radio receivers, forming one of the "plates" of the aerial-earth capacitance. Earthing also protects radio receivers from discharges of atmospheric electricity.

In the first case a special wire — the counterpoise — may be used instead of earth. But in the second case nothing can be used in place of the actual earth connection. Aerials frequently accumulate strong electrical charges, e.g., when a charged thunder cloud passes over an aerial or when a lightning strikes near by. Sometimes dry electrified snow or dust, raised by the wind, may charge aerials to a considerable potential. If the aerial is insulated from earth, its charge can flash over to the earth, damaging the radio receiver or causing fire.

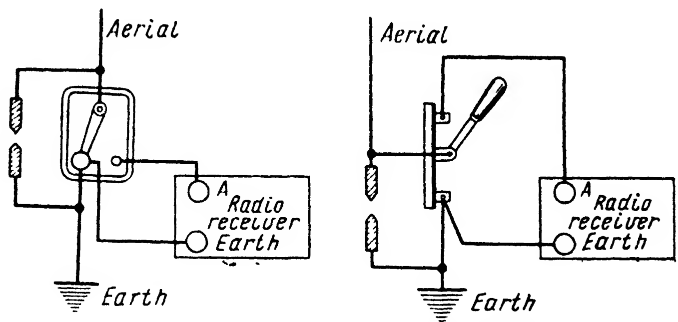


Fig. 46. Methods of connecting an earthing switch and a lightning arrester

Consequently receiving radio stations with external aerials *must have their aerials earthed when a thunderstorm is approaching and when no communication is being carried on.*

Single-pole double-throw earthing switches (two design versions of which are shown in Fig. 46) perform rapid switching of aerials from receivers to earth and back.

Such a switch should be installed on the window frame or near it, so that the aerial and earthing lead-ins may reach the switch by the shortest possible route.

An earthing switch should be always supplied with a lightning arrester, which is simply a 0.5 mm spark gap formed by two sharp pins or toothed plates connected to the aerial and earth. A lightning arrester provides additional safety; if the aerial is charged but remains connected to the receiver, the charge will leak off to earth in the form of small sparks through the 0.5 mm gap of the lightning arrester.

Many people mistakenly believe that aerials "attract" lightning. Lightning does not always strike the tallest objects, because it usually travels not in a straight line but in irregular paths determined by atmospheric parts with the smallest electric break-down resistance. It is impossible to tell in advance where the lightning will strike, for this depends upon the condition of the atmosphere at the given moment. Only on rare occasions does a lightning strike an aerial, and if the latter is earthed the damage will be smaller than in a building supplied with no such aerial or earthing facility.

In cities the earthing of radio equipment is usually achieved by connecting the equipment to water mains. When an independent earthing is required, it is secured by burying some metal object or driving a piece of water pipe into the ground. A wire connected to such an object or pipe will provide the required earthing.

Not every earth connection is good enough for radio reception. Earthing sunk in dry soil is often unsuitable because of its extremely high resistance. In cities an earthing sometimes conveys interference to radio receivers. Such interference may be caused by "wandering currents" usually generated by tramcars, which draw their current not only from the rails but also via all the shortest paths from the electric power station.

When a good earthing connection is not available, a counterpoise can be rigged up instead and connected to the radio receiver. This counterpoise is simply a wire suspended under the aerial and insulated from the earth.

Mobile radio stations nearly always use a counterpoise, because such stations constantly change their location and this makes it practically impossible to secure a good earthing connection at every site of operation.

Aircraft radio stations use the plane fuselage as a counterpoise. In such installations the radio station aerial is suspended by one end from the aircraft, the other end carrying a weight and dangling under the fuselage. This type of aerial is lowered by the plane after the take-off and rolled in before landing. In addition to the dangling aerial some planes also use a fixed (rigid) aerial stretched above and along the fuselage.

Sea-going vessels and river boats utilise a good natural "earth"— the water which they ply.

24. TRANSMITTING AERIALS

Transmitting aerials must be designed and built in such a way that the electrical power losses taking place in them are kept to the smallest possible value. This is important, because any reduction of useful power in the aerial of a transmitting station decreases the operating range of the station, which—if the losses in its aerial are high—will then fail to deliver a sufficiently strong signal to the aerials of distant receivers.

All energy losses in an aerial (losses in the ohmic resistance of wire, losses in insulators, etc.) can be conventionally considered as losses occurring in some resistance, which is called the *loss resistance*. The value of loss resistance in different aerials ranges from several ohms to tens of ohms. The consumption of power required by the radiation is considered as useful energy loss in another conventional resistance, known to radio engineers as the *radiation resistance*. Losses in the radiation resistance, connected into the current loop, give heat losses equivalent to radiation losses of the aerial.

If the radiation resistance R_{rad} and current I_A in the current loop of an aerial are known, the power of the waves radiated by the aerial can be determined by the following equation: $P_{rad} = I_A^2 R_{rad}$. The radiation resistance of a dipole operating on its fundamental frequency is 73 ohms.

Transmitting aerials are always tuned to resonance with the frequency of the transmitter, which secures the maximum aerial power, i.e., the maximum power of the waves radiated by the transmitter. Accordingly, the length of aerial wire of transmitters is usually determined by the wavelength range on which the transmitter operates. In aerial design the voltage developed in the aerial wire is an important factor; the higher the transmitter power, the higher is this voltage and the better should the aerial be insulated from earth.

Transmitting aerials of high-power transmitters have very large dimensions. They are of great height and are frequently designed for directional transmission, radiating their waves only over a single required sector.

Transmitting aerials built for omnidirectional transmission (transmission in all directions) are generally of the same configuration as the L-type, T-type, vertical and tilted types of aerials already studied. The length of the aerial wire is determined by the wavelength range of the transmitter, as mentioned above. Short-wave aerials are of comparatively small dimensions. Long-wave aerials, whose capacitance must be large, are built as high as possible. The horizontal part of such an aerial is frequently comprised of several wires, to increase the value of capacitance. Sometimes these wires are bunched to form a cylinder. Such aerials are often used

on ships. Transmitting aerials seldom employ earthing systems because of the losses, preference being given to various types of counterpoises.

Let us now examine several special types of aerials successfully used both for transmission and reception by short-wave and ultra-short-wave radio stations of low and medium power.

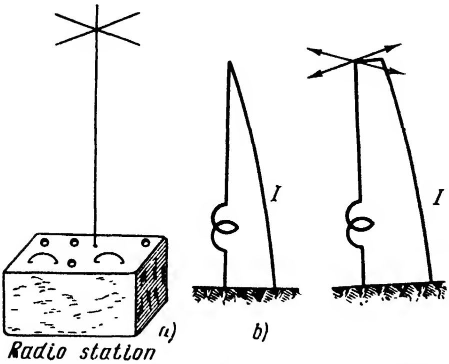


Fig. 47. A rod aerial and its current distribution

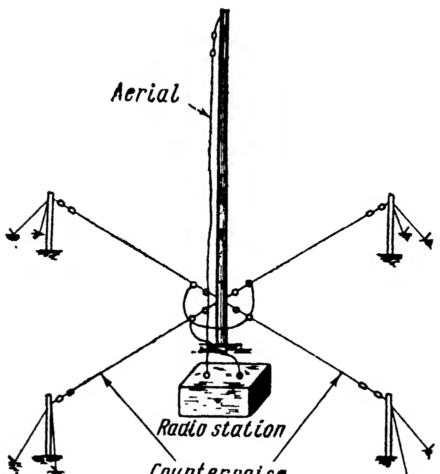


Fig. 48. A vertical aerial

Rod and vertical aerials. Portable and mobile radio stations, designed for operation over short distances, employ rod-type aerials. Such an aerial is a sectional metal rod, which is usually installed directly on the radio station case (Fig. 47*a*). The height of such a rod reaches 4 metres. The operating radius provided by the aerial can be increased by fixing a metal cross or brush at the tip of the rod. This slightly increases the aerial capacitance and changes the distribution of current in the rod. When no such cross or brush is attached to the rod, there will be a current node ($I = 0$) —and hence no radiation— at the tip; the metal cross part displaces the current node to its own tips and, consequently, there will be current and radiation at the tip of the rod (Fig. 47*b*). The metal case of the radio station is generally used as a counterpoise.

Greater radius of operation is offered by the vertical aerial. Constructionally such an aerial may be represented by a metal mast (a long rod) having a height of 6 metres or more. Alternatively, a wire suspended vertically on a wooden pole may be used (Fig. 48). Vertical aerials of this type are usually used in conjunction with a counterpoise consisting of several wires strung low over the ground. Neither the rod aerial nor the vertical aerial possess any directional properties.

Dipoles. Dipole aerials are frequently used by low-power portable radio stations. A dipole aerial, or simply a dipole, is made of two equal-length wires stretched out in a straight line (Fig. 49). A low-slung dipole gives the best radiation and the best reception in the directions in which the wires point. When communicating over short distances, the dipole wires, if insulated, may even be stretched out right on the ground. The two wire sections of a dipole may be also used in a different fashion, one section acting as a counterpoise and the other as an aerial. In such an arrangement the best operating range is obtained by lifting the free end of the "aerial section"; the maximum radiation will then take place in the direction in which the "counterpoise section" is pointing.

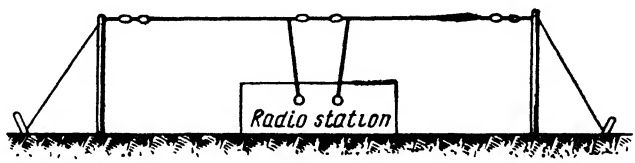


Fig. 49. A dipole aerial

When the operating range is great, the dipole is strung high over the ground and is fed with high-frequency power from the transmitter over special wires, which do not radiate and are called *feeders*.

Feeders with standing waves are always of two-wire type. In such feeders, as in an aerial, voltage and current distribution with attending loops and nodes takes place. The standing-wave feeders do not radiate because the currents in the two wires of a feeder flow in opposite directions with a 180° phase shift; if the feeder wires are placed sufficiently close to each other, this cancels the magnetic fields set up by the two currents.

In practice the distance between the two feeder wires is set to about 10-20 centimetres and is kept constant along the entire feeder length with the help of insulating spreaders, thus precluding any changes in the capacitance of the feeder. On ultra-short waves coaxial cables are usually employed as feeders. A coaxial cable is a tube-like flexible conductor, inside which another conductor is installed, insulated from the external one.

A standing-wave feeder must have a definite length, determined by the wavelength on which the aerial system operates. This is a disadvantage, as the length of the feeder has to be changed when the radio station is to operate on a different wave. Within certain limits such feeders can be retuned to different waves by means of shortening capacitors and loading coils, but this complicates the aerial system. Besides, a standing-wave feeder has high

losses, and its maximum length should not exceed 20-30 metres. Fig. 50 shows an aerial employing this type of feeder.

Another type of feeder, called *the travelling-wave feeder*, has practically supplanted the standing-wave feeder. Travelling-wave feeders can be of single-wire and double-wire types. They have no standing waves (no current or voltage loops and nodes). The length of a travelling-wave feeder is not critical, and since the losses in such feeders are low, feeder lengths of 100-200 metres are quite permissible. Fig. 51 illustrates aerials employing single-wire and double-wire travelling-wave feeders. Fig. 51 also gives the proper aerial dimensions for operation on the fundamental frequency. These aerials can be also excited on certain harmonics.

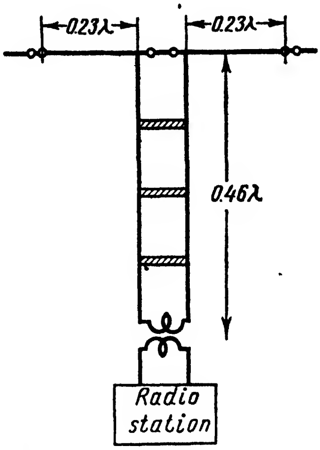


Fig. 50. A dipole aerial with a standing-wave feeder

The half-wave horizontal part of these aerial systems should be suspended high above the ground, and its directivity differs from that of a low slung dipole. The aerials studied here give maximum radiation in directions perpendicular to the aerial wire.

When highly directional transmission is desired complex aerial systems, comprised of many dipoles, are used in conjunction with so-called reflectors. On short and ultra-short waves these reflectors are additional dipoles located at definite distances from the

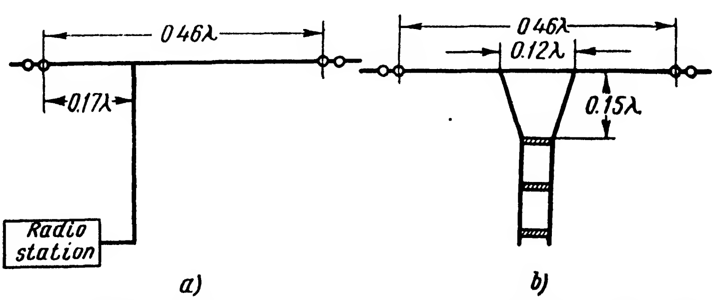


Fig. 51. Dipole aerials with single-wire and double-wire travelling-wave feeders

radiating dipole. On decimetric and centimetric waves the reflectors frequently consist of curved surfaces made of metal net and resembling the reflector of a searchlight or automobile headlight. A dipole (whose dimensions are very small on these extremely short

waves) is placed in the focus of the reflector. The *waves radiated by the dipole are gathered into a very narrow beam by the reflector*, and this beam — like the beam of a searchlight — can be pointed with great precision at any required target.

25. PROPAGATION OF RADIO WAVES

Before commencing the study of complex peculiarities of radio waves propagation, one should have a clear understanding of the atmosphere.

The air, as we know, consists of nitrogen, oxygen, hydrogen, and some other gases. Maximum density of the air is observed at the earth surface, where the air acts as a good dielectric. At higher altitudes the density decreases and the air becomes very rarefied. Nevertheless, the atmosphere boundary is located at very high altitudes, 1,000 kilometres and even higher.

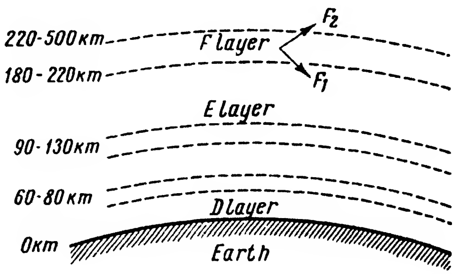


Fig. 52. Ionised layers of the atmosphere

In the *troposphere*, which is the lower layer of the atmosphere extending 10, to 14 kilometres up from the earth surface, all the gases are well mixed; within such altitudes the constitution of air is constant. But at higher altitudes (several hundred kilometres above the earth surface) the highly rarefied air begins to break up, and the gases of which it is comprised are distributed in layers, one above another, the relative position of each layer being determined by the weight of gas of which it consists. Thus at high altitudes the constitution of the atmosphere is not homogeneous.

The air becomes ionised under the influence of the sun's rays, cosmic rays and other factors, i.e., certain atoms of the gases constituting the air are broken up into free electrons and positive ions. Ionised air strongly influences the propagation of radio waves.

Maximum ionisation of various gases takes place at various altitudes, since different gas layers predominate at different altitudes. Investigation has revealed that the ionised layer of the atmosphere, the *ionosphere*, consists of several layers (Fig. 52).

D layer, existing only in the day-time, is located at an altitude of 60-80 kilometres; *E* layer — at 90-130 km; *F* layer — at 250-350 km at night. An interesting feature of the *F* layer is that in day-time it breaks up into two layers: F_1 lying at an altitude of 180-220 km, and F_2 — at 220-500 km.

Of course, there are no well-defined boundaries between these layers and the other parts of the atmosphere. The altitude, thickness and conductivity of the ionised layers are different at different times of day and year because of changes in the *ionising action* of the sun's rays. Moreover, the properties of the ionosphere change from year to year in an eleven-year cycle, which is also attributed to the activity of the sun. The higher the ionising intensity of the sun, the greater becomes the conductivity and thickness of the ionised layers, and the lower becomes their altitude. In the daytime their conductivity and thickness are greater and their altitude lower, as compared to night-time conditions. In summer their conductivity and thickness are greater and their altitude lower, as compared to winter.*

Every 11 years the maximum of the sun-spot cycle is repeated, these sun-spots being powerful sources of ionising emanations. This is when the conductivity and thickness of the ionised layers of the earth atmosphere reach their maximum values, while their altitudes decrease. Such are the complex laws governing the properties of the atmosphere, these properties governing, in their turn, the propagation of radio waves. Besides these periodic changes, there are also changes of chaotic character which cannot be foreseen.

Thus magnetic storms, which strongly affect radio reception, sometimes rage for many hours and even several days. Such magnetic storms are caused by powerful eruptions of electron streams by the sun, these streams reaching the earth atmosphere and strongly influencing the ionised layers. Layer F_2 is particularly affected by these phenomena; its conductivity decreases, its altitude increases, and the layer breaks up into clouds of electrons or else destroys itself completely.

When streams of meteors rush into the atmosphere at the altitude of E layer (about 100 km), this sometimes gives birth to the so-called *sporadic E layer*, noted for extremely strong ionisation, extending for not over one thousand kilometres and existing only for several hours.

Besides all these phenomena, chaotic fluctuations of ionisation continuously take place in the atmosphere, being stronger in the higher layers and particularly so in layer F_2.

All the above-mentioned changes, constantly occurring in the ionosphere, interfere with the normal propagation of radio waves and at times disrupt radio communication altogether. This makes it difficult to clarify the laws governing the propagation of radio waves and to calculate radio communication systems on the basis of such laws.

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* Only in F_2 layer the conductivity is greater in winter than in summer.

Important investigations directed at establishing a modern theory of radio wave propagation and calculation of radio communication have been conducted by *M. V. Shuleikin, L. I. Mandelstam*, N. D. Papaleksi, B. A. Vvedensky, V. A. Fok, A. N. Shchukin, A. G. Arenberg, and many other Soviet scientists.

In the propagation of radio waves over the earth surface in the atmosphere the following phenomena are observed.

Dissipation of radio wave energy. As a radio wave leaves the aerial of a transmitter and begins to travel in all radial directions from it, the wave energy is distributed over a constantly expanding space. Consequently, the amount of the wave energy in each point of the space gradually decreases.

Directional transmission, in which a radio wave is sent like a narrow searchlight beam to a required objective, is the only means of lessening the dissipation effect. Directional transmission increases the communication radius of a radio station and in many cases also achieves a high degree of secrecy, because the signal can be received only by receiving stations located in the path of the beam. Directional transmission is also employed by radio beacons, so important to aviation and sea-going ships, by radiolocation, the art of determining the position of various objects in any visibility, etc.

Radio wave absorption. The energy of radio waves passing through various substances is absorbed by such substances. It is only in interplanetary space that no absorption takes place. Non-ionised air absorbs little energy from the radio waves. Considerable energy is absorbed by solid dielectrics, semiconductors and conductors. When a radio wave meets any kind of conductor, the latter absorbs the greater part of the wave energy. This is only natural, as the wave creates a movement of electrons in the conductor and gives rise to a high-frequency current through it, and all this requires the energy which the conductor absorbs from the radio wave. On this phenomenon, incidentally, the principle of radio reception is founded.

On closer examination of radio wave absorption, we note the following. We already know that when a radio wave meets a conductor, i.e., runs into it, the conductor will absorb most of the wave energy. However, if the wave travels along the conductor, there will be much less energy absorption. This is why radio waves are propagated over much longer distances when they travel along highly-conductive surfaces, such as sea, rivers, railways and transmission lines. The propagation distance is considerably shortened when the wave is made to travel along poorly-conductive surfaces, such as stretches of dry land.

As mentioned above, dielectrics also absorb radio wave energy. The radio wave field causes displacement of electrons in the molecules of a dielectric — the displacement current. This is a high-fre-

quency alternating current, i.e., high-frequency oscillations of electrons inside the molecules. Displacement currents heat the dielectric, and this requires energy supplied by the wave.

Semiconductors combine the properties of conductors and dielectrics; both conductance currents and displacement currents are set up in them. This is why the ionised layers of the atmosphere, being semiconducting media, absorb to a noticeable extent the energy of radio waves passing through them.

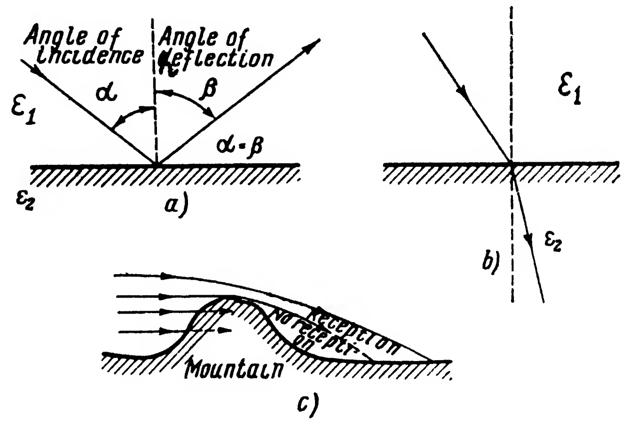

Fig. 53. Wave reflection *(a)*, refraction *(b)* and diffraction *(c)*

As the radio waves travel along the earth surface their energy is absorbed by the soil itself, as well as by local objects and obstructions, such as mountains, hills, woods, electrical transmission lines, etc. Metal roofs, reinforced-concrete buildings, wires, ore-containing mountains, damp earth layers, buildings laid with moist stone, forests, etc., absorb the energy of radio waves particularly well.

Reflection and refraction of radio waves. In a homogeneous medium radio waves travel in straight lines, but where a wave passes from one medium to another reflection and refraction take place. These phenomena occur at the boundary of two media with different dielectric permeabilities ε_1 and ε_2.

When a wave reaching the boundary between two such media turns back at a certain angle, reflection takes place (Fig. 53*a*). A wave reaching a flat surface at the right angle is reflected at the same angle, i.e., goes back the way it came. If a flat surface is reached by a parallel wave beam, the beam, after reflection, will remain parallel on its return from the surface. However, if the surface is uneven, the reflected waves will diverge. Conductors are the best reflectors of radio waves. The physical meaning of radio wave reflection is explained as follows. The arriving wave sets up currents

in the surface layer of a reflecting body, these currents producing the radiation of new electromagnetic waves, i.e., reflected waves.

When a radio wave passes from one dielectric to another, the wave is refracted, i.e., the direction of its travel changes (Fig. 53*b*). The greater the difference between the dielectric permeabilities ε_1 and ε_2 and the longer the wavelength, the more pronounced is the diffraction. The phenomenon of diffraction takes place because of the differences of speed with which radio waves pass through various substances.

Thus a radio wave which runs into a conductor is partly absorbed and partly reflected. Meeting a dielectric or a semiconductor, it is absorbed, reflected and refracted.

Diffraction of radio waves. The bending of waves around various obstacles is known as diffraction. If a radio wave meets some obstacle, such as a mountain, large building, etc., the wave is capable of going around it by bending its path (Fig. 53*c*). The longer the wave, the better is its capability of going around the obstacles. Of course, a wave can not turn very steeply. Diffraction is the reason why "zones of silence" are sometimes observed behind tall hills and metal structures. In such zones some radio stations cannot be received at all. However, the reception is restored a certain distance further along the path of the wave, also due to diffraction.

Wave interference. Wave interference is the intermingling of two or more radio waves in a given point of the space. The waves of various radio transmitting stations can intermingle, creating interfering squeals, howls, whistles, hum, and wheezes. If the interference is created between the sky wave and the ground wave of the same radio station, the level of the resultant signal will either rise or drop (sometimes to nil) owing to the phase difference of the two waves.

Waves radiated horizontally and propagated along the earth surface in the lower layer of the atmosphere are called *ground waves.* As they travel along the ground such waves are absorbed by it and also by various local objects. The higher the frequency of waves, the greater is such absorption. Depending upon their frequency, such waves bend around the earth curvature with greater or lesser ease, due to diffraction.

Waves radiated at an angle to the earth surface are called *sky waves.* Passing through weakly ionised parts of the atmosphere, the sky waves are only slightly absorbed and reach the ionosphere, where they are diffracted. Since the ionisation and dielectric permeability change gradually in the ionosphere layers, the path of a radio wave is represented by a smoothly changing curve. The longer the wave and the stronger the ionisation, the steeper is the bending of the wave. Fig. 54 shows layers E ans F_2 as observed at night. Beam *1*, corresponding to a short wave of a rather low frequency, strongly refracts in layer E and returns to earth. It is customary to

say in such a case that beam *1* is reflected by *E* layer. Beams *2* and *3*, corresponding to shorter waves, pass right through *E* layer because the degree of its ionisation is insufficient to turn these beams back. The ionisation of layer F_2 is insufficient to return beam *3* to the ground. There may be two reasons for this; either beam *3* is of too high a frequency, or else it enters *E* layer almost at the right angle and having reached the middle part of the layer, where ionisation is at its maximum intensity, does not turn sufficiently to reach the earth surface. Such a beam will then pass on into interplanetary space and is considered lost as far as radio communication is concerned.

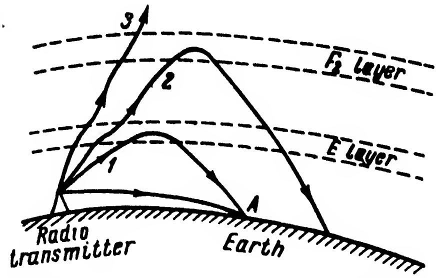

Fig. 54. The paths of radio waves in the atmosphere

The point at which beam *2* returns to earth is located at a greater distance from the transmitter than the point of beam *1* return.

In the ionised layers radio waves experience not only refraction but also absorption, the degree of which increases on longer waves. Since the altitude and intensity of the ionisation of ionosphere layers are changing, the paths of sky waves through the atmosphere change correspondingly. This explains the considerable variation of signal strength on short waves during the day and the year. It also explains the phenomenon known as *fading*. In most cases fading is caused by the arrival of several radio waves from the same transmitter to a receiving aerial, each having travelled by a different path. For instance, point *A* (Fig. 54) is reached by a ground wave and also by a sky wave, the latter reflected by *E* layer. Due to the changes constantly taking place in the ionosphere, the length of path of the sky waves is also changing all the time, which changes the wave phase. As a result of the adding of the ground wave and sky wave, the level of signal at the output of the radio receiver is constantly changing: when the phases of the two waves coincide, the signal becomes stronger; when the phases are opposite, the signal level drops (sometimes to nil).

It is rather difficult to find effective measures against fading. One of the most successful methods devised up to now is called diversity reception. In this, the receiving station employs two or three aerials, spaced at a distance of 200-300 metres from each other. The radio receiver connected with such aerials is provided with an individual high-frequency amplifier and a detector for each aerial, but employs a common low-frequency amplifier. The method of diversity reception is based on the assumption that the fading effect does not accur simultaneously at various points of topography.

While the signal is fading in one of the aerials, its strength is increasing in another. Thus the signals compensate each other, and the combined low-frequency signal at the output of the receiver changes only very slightly.

Automatic gain control circuits, used by most radio receivers, also help to some degree to counteract fading (see Chapter IX).

Let us now examine the propagation peculiarities of waves of different lengths. 10 - 100 KHz.

Long waves. Long ground waves (3,000-30,000 metres) propagate by curving around the earth surface and certain obstacles, which is possible owing to their considerable diffraction ability. However, the earth and obstacles absorb much of the energy of these waves. Sky waves of this wavelength range are reflected from the ionosphere (in day-time from D layer, at night from E layer), return to earth, reflect from its surface, return to the ionosphere, travel back to earth, etc. Strong absorption of wave energy occurs during these repeated reflections, and this is why long distance communication calls for high-power long-wave transmitters. Fading is completely absent on long-wave communication. In winter and also at night long wave reception is somewhat better than in summer and during the day, which is logical because the air is less ionised in winter and at night, and hence the wave absorption is reduced. Various other changes in the ionosphere and troposphere do not practically effect the propagation of long waves. In comparison with waves of other ranges, long waves offer the most constant conditions for propagation.

Medium waves. A great deal of attention is paid to the medium waves (200-3,000 metres), because they are used for broadcasting. Sky waves of this range are very strongly absorbed by the ionosphere in day-time and are of no practical importance from the viewpoint of radio communication. Ground waves of the same range are also strongly absorbed by the earth. The shorter the wave and the worse the conductivity of the surface layer, the greater the absorption. Communication over sea water suffers the least absorption, and over dry land the strongest. Thus medium waves propagate over comparatively short distances in day-time and over much greater distances at night, because in the latter case the absorption of these waves is much smaller during their reflection from the ionosphere. The medium wave communication range is greater in winter also because of reduced ionosphere absorption.

During night reception of medium waves, considerable fading is often experienced. This is caused by the interfering sky and ground waves, which travel by various paths and reach receiving aerials out of phase. Other ionospheric changes exert hardly any influence on the propagation of medium waves. Strong atmospheric interference, created by various electrical discharges occurring in the atmosphere and particularly severe in summer, accompanies medium-wave reception.

It should be remembered that the conditions of propagation do not change abruptly but gradually from one wavelength range to the following one. Hence medium waves of the 2,000-3,000 metre range resemble the long waves, as far as their properties are concerned. The peculiarities of the medium waves, explained above, are most pronounced on 200-600-metre range.

Short and intermediate waves. It should be noted that short and intermediate waves (10-200 metres) are strongly absorbed by earth

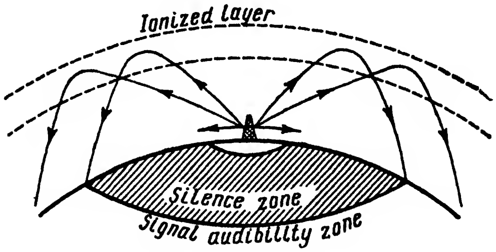

Fig. 55. Short-wave zone of silence

and local objects. Therefore the ground wave has a very short operating radius, which usually does not exceed several dozen kilometres. The less the energy and the shorter the wave, the shorter this distance. Beyond this distance, in waves shorter than 80 metres, begins *a zone of silence* (Fig. 55).

Depending upon the wavelength and upon the time of day and year, such a zone of silence can stretch from several hundred to several thousand kilometres. The shorter the wave, the greater the zone. It is greater at night than in day-time, and greater in winter than in summer.

Beyond the zone of silence begins a zone of audibility, i.e., a zone where the signals of the given station are heard once again. These signals, however, are not delivered to the remote audibility zone by the ground wave (which was completely absorbed a long way back). The signals now arrive via the sky wave path, the sky wave having been reflected from the ionosphere back to earth. Reception is quite good in this zone, although the signals occasionally fade. Such fading is nearly always present on short waves and at times is very sharp and frequent.

As a rule the zone of silence is absent on waves of 80-200 metres but is sometimes encountered at night on 50-80-metre waves. Waves of 35-70-metre (and longer) are used chiefly for long-distance night communication and are, accordingly, called *night waves.* These waves can be also used for day-time communication, but only over short distances (ground wave communication).

On short waves the sky wave is usually reflected from F_2 layer and loses some of its energy by absorption in layer E, twice passed

by the wave. In long-distance day-time communication night waves give weak reception because they are strongly absorbed by *E* layer. Their performance is improved at night because of reduced absorption in *E* layer at this time.

Waves of the 10-25-metre range are only slightly absorbed by *E* layer and are therefore used for day-time communication and called *day waves*. On these waves the silence zone is considerably larger, particularly at night. Waves of 10-25-metre range are not very suitable for night communication, because at night the ionisation of F_2 layer is insufficient to turn them back to earth. When long-distance operation is required, waves of the 25-35-metre range are employed both at night and in day-time. Radio broadcasting stations, for instance, chiefly use the 10-35-metre range for day operation and switch over to the 25-70-metre range at night.

Short waves cannot ensure constancy of communication, and the boundaries of their silence zones change because of altitude and conductivity changes of ionosphere layers during the day, during the year and during the 11-year sun activity cycle. Waves of different lengths have to be employed for different periods of day and night, summer and winter. Nevertheless, short waves possess a tremendous advantage over the waves of other ranges, i.e., the possibility of communication over thousands of kilometres with transmitters whose power often does not exceed several watts. Cases are known when communication has been established half-way around the globe (20,000 kilometres) with such low-power short-wave transmitters.

Various agitations (e.g., magnetic storms) taking place in the ionosphere strongly influence the propagation of short waves and sometimes make communication on these waves impossible. Such disruptions of short-wave communication are most frequently observed in the regions located close to the earth's magnetic poles, because to such regions are directed the fluxes of electrically charged particles entering the earth atmosphere from the sun. The duration of such disruptions of short-wave communication usually does not exceed several hours. When such disruptions occur, communication has to be maintained on longer waves or through relay stations located at larger distances from the magnetic poles of the earth. The present state of radio technique makes it possible to predict ionospheric agitations with considerable accuracy, because such agitations are related to the periodic changes of the sun's activity and, partly, to the rotation of the sun on its axis.

Information concerning forthcoming ionospheric agitations and the character of changes in the conditions determining the propagation of short waves is published in the so-called radio forecasts. Apart from such periodic agitations, non-periodic ionospheric agitations are also observed but cannot be predicted.

A very interesting phenomenon of radio-echo is also observed *on short waves*. *Here the signals of a radio transmitting station* reach a radio receiver over a dual path, one path being the shortest distance—say, 2,000 kilometres—between the sending and receiving points, and the other path passing right around the globe, which, in the given example, constitutes 38,000 kilometres.

Signals following the second path, quite understandably, reach the receiver with some delay (one-seventh of a second in the given case). The passing of short waves around the globe is explained by their repeated reflections from the ionosphere and the earth surface. In this way a short wave can even circle the globe twice, as shown by corresponding radio-echo investigations.

Short waves have another important advantage over waves of other ranges. They are more immune to various types of atmospheric and industrial interference, and communication on these waves can be carried on despite electrical noise produced by tram-cars, electric welding, etc., when the reception of longer waves is impossible. The shorter a wave, the more immune is its reception to such interference.*

Metre, decimetre and centimetre waves. As a rule waves shorter than ten metres are not reflected by the ionosphere and, passing right through it, go off into interplanetary space. Cases are known when such waves have been reflected from the Moon and even from Venus. On these extremely high frequencies only the ground wave is usually employed for practical radio communication purposes. However, these waves are strongly absorbed by various local objects and, moreover, are hardly able to diffract. It is therefore important that there are no obstacles on the optical line between the transmitting and receiving aerials operating on such waves. When the distance between the aerials is sufficiently large (several dozen kilometres) for the curvature of the globe to be taken into consideration, the aerials should be elevated to a great height. On these waves low aerials can maintain communication only over short distances of several kilometres (the sight-line distance). Various local objects, such as buildings, woods, etc., represent serious obstacles to the given type of communication. Ultra-short-wave television transmissions are sometimes picked up at a distance of over 100 kilometres when the aerials of TV transmitting stations are highly elevated.

Thus for ground radio stations *the waves of the metre, decimetre and centimetre ranges are chiefly used for communication over fairly short distances, whereas communication with spaceships on ultra-short waves is carried out at enormous distances of hundreds of thousands and even millions of kilometres.* The main advantages of ultra-short

* Certain types of interference, for instance, automobile ignition noise, are more bothersome on short waves than on waves of other ranges. More information on interference is given in Chapter IX.

waves is their small degree of fading and independence of their propagation upon the time of day and year, which is easy to understand since the propagation of such waves is not influenced by the ionosphere. Waves of these ranges also possess other advantages, i.e., ease of directional transmissions and almost complete absence of interference.

In practice ultra-short-wave stations communicate over distances slightly exceeding the line-of-sight ranges when the longer waves of the ultra-short-wave range are employed. This is achieved because of the slight bending of such waves around the earth's curvature, but chiefly because of refraction in the troposphere. The refraction, in this case, is attributed to the fact that various air layers have different density, temperature and quantities of water vapour, which accounts for their varying dielectric permeability. Since the condition of the air close to the earth surface is constantly changing, the refraction of ultra-short waves vary correspondingly and beyond-the-horizon communication on these waves is not very stable. The important advantage of ultra-short-wave communication—the stability of reception—is usually attained only over line-of-sight distances.

Owing to the development of television and ultra-short-wave amateur communication, numerous cases have been recorded when ultra-short-wave transmissions were picked up at distances of several hundred and even thousand kilometres from the transmitting stations. Usually such reception is irregular and is accompanied by the fading of signals.

There are several reasons for such freakish super-long contacts on ultra-short waves. First of all, in years when the sun activity is at its maximum, the ionosphere sometimes reflects waves as short as 6-7 metres in day-time. Moreover, the creation of fairly short-lived (sporadic) strongly ionised layers of the ionosphere — electron clouds — results in the reflection of waves whose lengths may be as short as 3 metres. In particular meteorites which enter the earth atmosphere create ionised areas (meteorite trails) in their paths. Another reason for the super-long reception of ultra-short waves is the changes of temperature and moisture content in the troposphere, which in some cases causes intensified atmospheric refraction of ultra-short waves. This creates the possibility of ultra-short-wave propagation via their repeated reflections from some layer of the troposphere and the earth. Finally, ultra-short waves may at times be reflected by certain tropospheric irregularities existing at various altitudes. In recent years such reflection has been successfully employed for the establishing of regular and dependable communication on ultra-short waves over great distances. This is possible because of the great sensitivity of modern receiving apparatus.

It should be noted that waves of the centimetric range are subjected to considerable absorption in the atmosphere. Thus waves

shorter than 5 cm are absorbed in the air by various drop formations, such as *rain, fog* and *snow*. A wave of 1.3 cm is even absorbed by water vapours. All this creates difficulties in the application of extremely short waves.

26. QUESTIONS AND PROBLEMS

1. Why does an open oscillatory circuit radiate electromagnetic waves much better than a closed one?
2. Why is the aerial of a radio transmitter tuned to resonance?
3. A single-wire L-type aerial is comprised of a 30-metre horizontal part and a lead-in (vertical part) 5 metres long. Determine the inductance and capacitance of the aerial.
4. The full length of the wire of an unearthed aerial is 20 metres. Find the natural frequency and natural wavelength of the aerial.
5. How can the natural wavelength of an aerial be decreased or increased?
6. What are the properties of a loop aerial?
7. If a charged capacitor is placed between the poles of a magnet, it will be acted upon both by the electric and magnetic fields. In this case, would it be correct to say that the field inside the capacitor is electromagnetic?
8. What are the units of measurement for the field intensity of radio waves?
9. Is it a good practice to install an aerial under an iron roof in a garret?
10. Why is it that at night the ionised layers of the atmosphere are at higher altitudes and of less ionisation than in day-time?
11. What is the diffraction of radio waves?
12. What are the propagation peculiarities of ultra-short waves?
13. If in the opposite ends of a house are located two radio receivers, it is possible to suspend the two required L-type aerials with the help of two aerial masts instead of four. How would you do it?
14. Devise a circuit diagram for earthing an aerial with an usual type of switch in lieu of a lightning arrester.
15. What is the function of the horizontal part of a transmitting aerial?

CHAPTER IV

ELECTRON VALVES

27. DESIGN AND OPERATING PRINCIPLE OF THE TWO-ELECTRODE VALVE

Electron valves are the most important parts of radio transmitters, receivers, amplifiers and rectifiers. They can generate, detect, amplify and rectify alternating currents of various frequencies and have found wide application in many other branches of modern science and engineering apart from radio.

In Russia the first electron valves were made in 1909-10 by V. I. Kovalenkov, who used them for the amplification of audio-frequency currents. Later, in 1914-15, several types of valves were designed and built by M. A. Bonch-Bruyevich, and in 1919 he published the theory of the three-electrode valve. The Svetlana Plant has played an important role in the production of Soviet radio valves.

The action of electron valves is based on the flow of streams of free electrons through a vacuum. Hence every electron valve must be provided with means of obtaining sufficient quantities of free electrons. The phenomenon of emanation of free electrons by the surface of a substance is called *electron emission*.

When the electrons are emitted as a result of thermal action, such emission is known as *thermionic emission*.

Other types of emission are as follows: *electrostatic* or *autoelectronic emission* — forcing the electrons out of a substance with the help of a powerful external electric field; *secondary electron emission* — knocking the electrons out of a substance with the help of fast moving electrons; *ion-activated emission* — knocking the electrons out with the help of ion blows; *photoelectron emission* — emission of electrons under the influence of light beams.

Most electron valves operate by thermionic emission, based on the following phenomenon. When a conductor is heated to a high temperature, it begins to give off free electrons into the surrounding space. Prior to their liberation these electrons remain in the conductor, moving in it in a chaotic way. The speed of the electrons increases upon heating, and when the temperature is sufficiently high

they move at such a great velocity that some of them fly out of the conductor.

A *diode*, or two-element valve, is the simplest of all electron valve types. The two electrodes of which it is comprised are located inside a glass or metal container with a high vacuum (Fig. 56*a*). *A filament,* othervise referred to as *a cathode*, represents one of the electrodes. The other electrode — *the anode* — may be represented by a metal plate. Fig. 56 shows the symbol indicating diode valves in circuit diagrams. In symbolic representation the filament is shown in Fig. 56*b* by an arc or a triangle, and the anode by a dash. Two wires are connected to the filament, and one to the anode.

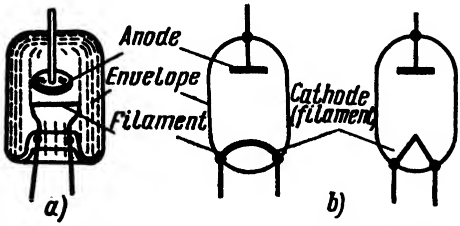

Fig. 56. Design and schematic representation of a diode

The electrons are emitted by the cathode (filament). The electrons given off every second constitute the emission current, or simply emission, which is usually expressed in milliamperes.

When the temperature is low the emission is practically absent. With a rise of temperature the emission increases, at first slowly and then faster, reaching a considerable value at temperatures of several hundred degrees and higher. It is not advisable to increase the temperature too much, seeking higher values of emission, because in the end the filament will become overheated and will melt or, as is usually but not quite accurately said, burn out.

Thus, *the higher the temperature, the greater the emission.*

Increasing the surface of the cathode increases the emission, too. The value of emission is also determined by the material of which the cathode is made.

The function of the anode in an electron valve is that of attracting the electrons emitted by the cathode and thus creating a flow of free electrons in the valve.

The anode must be charged positively in order to attract electrons. The existence of electrical field between the anode and cathode accounts for the attraction of electrons by the positively-charged anode, for it is under the action of this field that the electrons given off by the heated cathode move to the anode (Fig. 57). The container, also called envelope, ensures that the described process takes

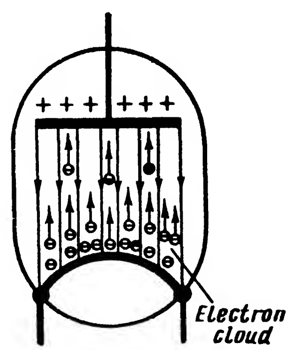

Fig. 57. The action of the anode electric field upon the electrons in a diode

place in vacuum, which is absolutely necessary for the following reasons.

The air is pumped out of the envelope in order not to interfere with the flow of electrons between the cathode and anode. Further, if air were present in the envelope, the heated filament would burn out due to chemical reaction between the metal and oxygen. It follows that the normal operation of an electron valve requires a very high vacuum in its envelope.

If the vacuum is insufficient, the electrons will be striking air molecules during their flight from cathode to anode, thus causing their ionisation. These molecules, having some of their electrons knocked out by the collisions, will be transformed into positive ions. The ions, being repelled by the positively-charged anode, start moving to the negative cathode, creating a so-called gas current and thus upsetting the normal operation of the valve. An electron valve with traces of air ("soft valve") does not function properly. A good valve should contain not more than a 10^{-9} part of the air it contained before evacuation.*

It is rather difficult to obtain a sufficiently high vacuum in an electron valve. The air is first pumped out by means of primary pumps, after which high-vacuum pumps are employed. But even this is insufficient; therefore when the valve is being made, a small piece of magnesium or barium, called the getter, is placed in it. After the pumping is completed and there are still considerable traces of air in the valve, the latter is subjected to heating, causing the getter to evaporate. Upon cooling, the getter material is deposited on the inside walls of the envelope, giving the latter a characteristic hue (mirror-like surface for megnesium, brownish-black for barium). This metal deposit absorbs the last traces of air and other gases given off by the electrodes of the valve during normal operation in a radio circuit. Such is the procedure of creating and maintaining high vacuum in an electron valve.

To provide a convenient way of connecting a valve to an electrical circuit the valve envelope is fixed to an insulated base supplied with embedded metal pins. To these pins are soldered the leads from valve electrodes. In many types of valves the pins are embedded directly into the glass envelope of the valve. Such designs offer a simple and practical construction; it is only necessary to insert the valve pins into a specially designed socket, and the desired electrical connection of the valve elements to the rest of the circuit is automatically made.

Now a few preliminary words about the construction of the valve elements. In valves with cylindrical construction of electrodes

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* Special valves, called ionic or gas-discharge devices and making use of the ionisation phenomenon, operate properly only when filled with gas. (See Sections 51 and 58.)

(Fig. 58*a*) the anode is shaped as a tube, while the filament is stretched into a single straight line or else double back, resembling letter V. In valves with rectangular electrodes the anode has the shape of a box, within which is located a V-shaped or M-shaped cathode (Fig. 58*b*). There are also other configurations of electrodes. The anode is generally made of some high-heat metal, such as nickel, molybdenum, or tantalum, though some valves employ anodes made of graphite. The cathode is made of tungsten or some other metals. The construction of cathodes is described in Section 31.

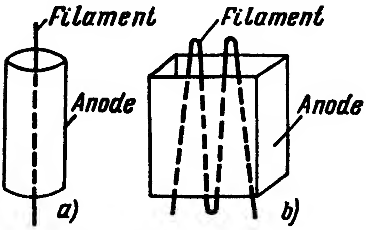

Fig. 58. Design of diode electrodes

28. DIODE CIRCUITS

Fig. 59 gives a schematic representation of diode circuits.

The battery used to heat the cathode is called the filament battery, or LT (low-tension) battery. The circuit comprised of this battery and the filament of the valve is called the filament circuit.

The filament is designated by letter F and the cathode in indirectly-heated valves (explained later) by letter C. Current flowing through the filament is denoted by I_f, and the filament voltage, i.e., the voltage applied to the filament ends, by U_f. Filament

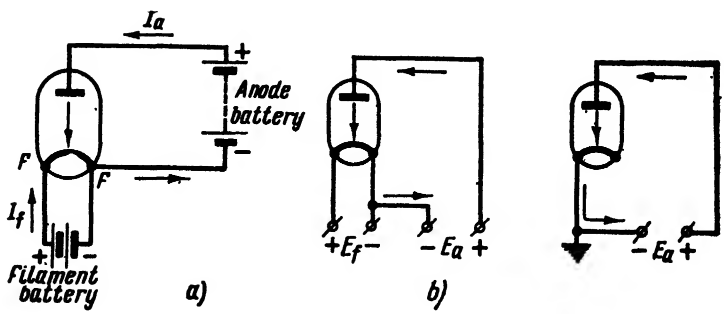

Fig. 59. Various methods of schematic representation of diode circuits

voltage is measured by a voltmeter connected in parallel with filament ends, the value of this voltage being adjustable by means of a rheostat connected in series with one of the filament terminals (the voltmeter and the rheostat are not necessary in practical radio circuits, where the value of filament voltage is fixed and remains fixed during operation).

Filament voltage is always low and does not exceed several volts in low-power valves. In such valves the filament current equals several tens or hundreds of milliamperes. When several valves are used in a radio set, their filaments are connected in series if the voltage of the *LT* battary is approximately equal to the normal voltage of an individual valve. If the voltage of the *LT* battery is considerably greater than such a value, the filaments of the valves are connected in series (provided they are designed for equal values of current); as an alternative, they can be connected in series-parallel. If the voltage of the *LT* battery is excessive, it may be lowered with the help of a rheostat or a fixed resistor, connected in series with the filament circuit.

The battery connected between the cathode (filament) and the anode is called the anode battery or HT (high-tension) battery. The circuit comprised of this battery and the space between the anode and cathode (inside the valve) is referred to as the anode circuit. All values pertaining to this circuit are designated by letter *a*.

The current flowing through the anode circuit is called the anode current and is designated I_a. *This current is, in effect, the stream of electrons flowing from the cathode to the anode inside the valve and on to the external part of the anode circuit.*

Electrical engineers traditionally consider that electrical current flows from the positive to the negative terminal of the power supply, feeding an external load. This is indicated by arrows in Fig. 59. However, when studying electron valve circuits the direction of electrical current flow, i.e., the true direction of flow of electrons, is assumed to be from the negative to the positive terminal of the power supply when the current flows through a valve circuit.

The anode current is of greatest significance in an electron valve circuit. Starting from the negative terminal of the *HT* battery, this current flows as follows. First it reaches the cathode, then the latter starts emitting electrons, which fly to the anode, after which the anode current path is completed to the positive terminal of the *HT* battery. The anode current can flow only when the filament voltage is sufficient, when the anode is positive in relation to the cathode, and when the anode circuit is complete.

The potential difference between the anode and cathode is called anode voltage, denoted by U_a. This is a very important factor. In the circuit given in Fig. 59 the anode voltage is equal to the voltage across the terminals of the *HT* battery.

When studying processes taking place in any electronic equipment it is customary to regard the cathode potential as zero in respect to the potentials at all other electrodes of an electron valve.

Always keep in mind that it is the anode voltage which makes the anode current flow. *Heating the cathode is the function of the filament circuit. Producing the anode current flow, when the thermionic emission is sufficient, is the function of the anode circuit.*

In the circuit of Fig. 59*a* one end of the filament is connected to *the negative terminal of the LT* battery *and also to the negative terminal of the HT* battery. This point of connection is called *the common negative* and is usually also connected to earth and to the metal chassis of a radio set. All the voltages of the electrical circuit are then measured in relation to this point.

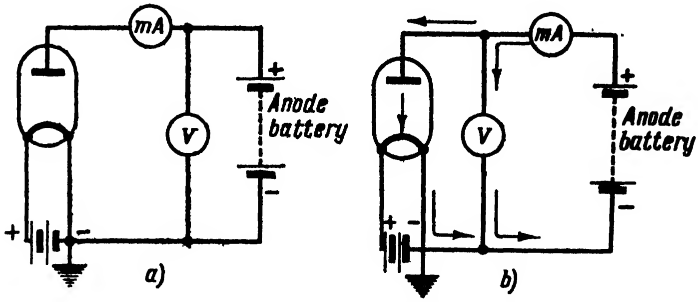

Fig. 60. Measurement of anode current and anode voltage: *a)* correct connection of measuring instruments; *b)* incorrect connection of measuring instruments

In some sets the negative terminals of *LT* and *HT* batteries are connected to each other right at the place where the batteries are installed. In such cases the common negative wire running to the set carries both the filament and the anode currents.

Electron valve circuits can have different schematic representations. Thus in Fig. 59*a* the *LT* and *HT* batteries are actually shown with respective symbols, while Fig. 59*b* shows only the terminals to which these batteries are connected. For the sake of simplicity the circuit carrying the filament current is often not shown in full. In this case a circuit diagram shows only that filament wire which is connected to the negative side of the *HT* battery (the common negative), as represented in Fig. 59*c*. Sometimes a circuit diagram shows only one positive terminal of the *HT* battery; in such cases it must be understood that the negative terminal of the *HT* battery is connected to the chassis of the radio set. In future we shall represent the source of anode supply in different ways, because the anode voltage may be supplied not only by a battery but also by other devices and apparatus, e.g., by a rectifier.

The anode voltage used by small valves can have a value of several hundred volts. The anode current is always smaller than the filament current and is equal to several milliamperes or several tens of milliamperes in low-power valves. A milliammeter connected into the anode circuit is used to measure the anode current. The anode voltage is measured with a voltmeter, connected across the anode and cathode (Fig. 60*a*). Fig. 60*b* shows how the two measuring instruments should NOT be connected, because in such an arrangement

the milliammeter will show the total current consumed by the anode circuit and by the voltmeter.

A diode valve passes the anode current in one direction only. The electrons can move only from a negatively charged cathode to a positively charged anode. If the anode is charged negatively and the cathode positively, no anode current will flow in the valve. A negatively charged anode repels electrons.

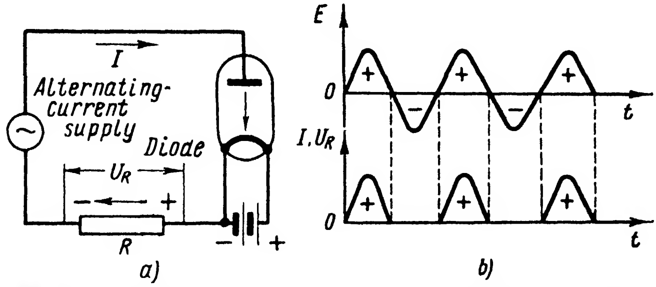

Fig. 61. Circuit diagram and graphic representation of alternating-current rectification by means of a diode

Thus a diode valve can conduct anode current in one direction only, which accounts for the name — valve. This property of a diode makes it useful as a rectifier, converting an alternating current into direct current, i.e., into a current which flows in one direction.

The fundamental circuit of a diode rectifier is shown in Fig. 61*a*. The circuit is comprised of an alternating-current generator, diode and load resistor *R*, all three being connected in series. The generator produces alternating e.m.f. (Fig. 61*b*), but the current in the circuit and the voltage across the load resistor will have a pulsating character (Fig. 61*c*) due to the rectifying action of the diode. In such a circuit the diode functions as follows: when the polarity of voltage developed by the generator is such that the anode is charged positively in respect to cathode, the diode will pass current (positive half-wave) and voltage will appear across the load resistor. During the following half-wave the polarity will be reversed; the anode will be charged negatively, the cathode positively, the diode will not pass current, and there will be no voltage across the load resistor. Hence the pulsating character of the current flowing through the circuit. Note that in the described circuit only positive potential will be applied to that end of the load resistor which is connected to the cathode of the valve, which accounts for the fact that when the diode is conducting current, it will always flow in one direction only.

In some cases this rectifier circuit is modified, but its general operating principle remains unchanged wherever diode valves are employed for the purpose of rectification.

29. TYPES OF CATHODE

In many valves the cathodes (filaments) are made of high-heat tungsten, which does not melt at temperatures below 3400° C. Valve filaments were formerly made of pure tungsten, but this was found to be very uneconomical because such filaments had to be heated to very high temperatures and this required too much electrical power for this filament circuit.

Activated cathodes were invented later. These cathodes are made of tungsten or other metals whose surface is covered with active metals or oxides. Such active metals and oxides readily emit electrons even at comparatively low temperatures. Hence an economy of filament power and the reason why nearly all modern electron valves employ activated cathodes. Pure tungsten filaments are still used only in certain high-power transmitting valves.

When the filament temperature is raised, the cathode emission is increased, but the service life of the valve is lowered. Therefore cathodes are so designed that they produce maximum emission at minimum possible temperatures. This saves filament power and prolongs the service life of the cathode.

Let us now examine basic types of cathode.

A tungsten filament produces emission at a temperature of 2300° C, which corresponds to white or light-yellow heat. The emission from such a cathode is much smaller than that from an activated one, but a pure-tungsten cathode has its own advantages. It gives a constant value of emission throughout its service life and, besides, withstands overheating (overvoltage) well. Temporary overheating does not decrease the emission of a tungsten filament. At the same time, the emission of activated cathodes is not very constant, is easily lost on overheating, and, once lost, cannot be restored. Of course large overheating is dangerous also in the case of a pure-tungsten filament, because it will simply burn out.

Activated cathodes lose their emission on overheating owing to the evaporation of the active layer from their surfaces under the influence of abnormally high temperatures. This is why an activated cathode with lost emission still possesses a mechanically undamaged filament. The service life of an activated cathode is over when its emission decreases by 10-20% due to the used-up active layer. After a long period of operation the emission of a pure-tungsten cathode is also reduced owing to another factor; due to the constant influence of a very high temperature, a pure-tungsten filament gradually grows thinner and its emission decreases because of the reduced area of the emitting surface.

Another disadvantage of activated cathodes is their somewhat unstable operation on high anode voltages. This is explained as follows: even in a good electron valve there still remain slightest traces of air and under the influence of high anode voltage this leads to

the appearance of ions in the valve; such ions strike the negatively-charged cathode with sufficient force to destroy the active layer. Cathodes made of pure tungsten are obviously immune to this sort of ion bombardment.

Valves with pure-tungsten cathodes employ no getters, and therefore their glass envelopes are transparent. These valves need no getters because tungsten particles, evaporating during normal operation of the valve in radio equipment, are deposited on the inner walls of the valve envelope, and the deposited layer absorbs the gas traces.

At present three following types of activated cathode are employed in electron valves.

Carbidised cathode. This type is made of tungsten or molybdenum with impurities of metallic thorium and carbon. Formerly thoriated cathodes were used, but these cathodes easily lost their emission because they contained no carbon. Carbidised cathodes operate at a temperature of about 1700° C, and their emission is more stable. They are used in medium-power transmitting valves and can operate at anode voltages up to 1,500 volts.

Oxide-coated cathode. This type is made of nickel or platinum wire with a coat of metallic oxide, such as barium, strontium or calcium. The working temperature of such cathodes is about 800° C, and their emission is far greater than that of tungsten or carbidised cathodes. Oxide-coated cathodes are widely used in very many types of valve, but their emission is not very constant and the valves cannot continuously operate on high anode voltages. Such valves withstand short and slight overheating, but lowered filament voltage can decrease the emission or even cause the cathode to burn out due to the appearance of local overheating points in the oxide layer.

Recently oxide-coated cathodes have been successfully used in so-called pulse work. On short-duration pulses of anode and grid voltages such a cathode can give a much higher emission than on continuous operation.

However, in such a case the cathode must be given a rest so that the oxide coat may accumulate a sufficient quantity of electrons ready for emission during the next pulse.

Bariated cathode. These are made of tungsten wire coated with copper, the surface of which carries a layer of barium oxide or metallic barium. This type is no longer used in modern valves.

All valves employing activated cathodes are easily distinguished in operation by the weak glow of their filaments. Carbidised cathodes glow yellow, oxide-coated cathodes glow red, and bariated cathodes glow so weakly that the glow is hardly perceptible. Moreover, the getter material leaves a mirror-like or dark deposit on the glass envelopes of activated-cathode valves.

Thus far we have been interchanging two words "filament" and "cathode". The words, in both cases standing for the electron-

emitting electrode of a valve, are synonymous. But there is a difference *which calls for exact classification* as we advance in our studies *of electron valve* design.

The filament serving a dual purpose of heating and of direct electron emission will from now on be called simply a filament, which is the accepted practice. Accordingly, all valves using this type of emitter will be referred to as *filamentary valves*. Filamentary valves are used in battery radio receivers and in portable radio stations, where minimum consumption of filament-heating electrical

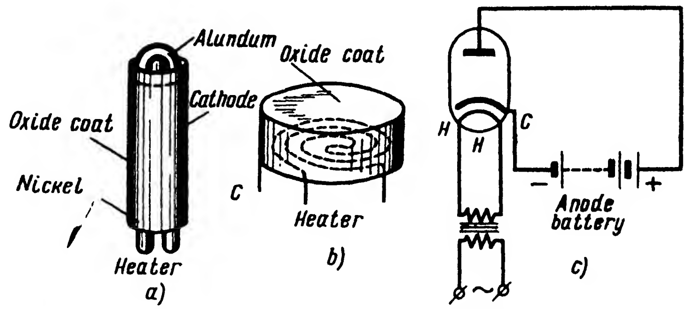

Fig. 62. Construction of indirectly-heated cathodes and a circuit diagram of a cathode-type valve

energy is a very important consideration. In most cases alternating current cannot be used for feeding the filaments of electron valves, because the heating of such filaments will periodically fluctuate at a speed equal to double the frequency of the feeding current. The emission of the filament would then be pulsating at the same speed, which is, of course, very undesirable. And only in special filamentary valves with thick filaments, do the latter possess sufficient thermal inertia to prevent fluctuation of the emission as the alternating heating current goes through its evolutions. There are very few types of filamentary valve with such thick filaments and, therefore the filament of a filamentary valve cannot as a rule operate on alternating current, which brings us to the subject of cathode-type valves.

Radio receivers and amplifiers designed for operation from alternating current mains use valves with so-called indirectly heated cathodes. A. A. Chernyshov was the first inventor to develop this type of valves, which are now known as *heater type valves*, *equipotential-cathode valves*, or simply *cathode-type valves*.

A modern cathode-type valve is shown in Fig. 62*a*. Unlike a filament, which serves the dual purpose of heating and emission, an indirectly-heated cathode serves only one purpose — that of emission. The heating in this case is done by another independent element, called the heater, placed next to the cathode.

In the type of valve here studied a cathode is usually represented by an oxide-coated nickel pipe into which the heater is inserted.

There is no electrical connection between the cathode and the heater because the latter is coated with special heat-resistant insulation. The heater, which is inserted into the cylindrical cathode, has the shape of a straight loop or a spiral, and its only function is that of heating the cathode which does the actual emission.

The thermal inertia of such a cathode is so high that it takes many seconds for it to heat up to the point of emission and the same length of time to cool off. Hence when the heater is supplied with power from alternating current mains, the cathode temperature stays constant. One disadvantage of an indirectly heated cathode is its slow start, i.e., 20-40 seconds of waiting before the emission begins.

One of the design versions is a cylindrical cathode with a bottom carrying an oxide-coated layer (Fig. 62*b*). A heater is also placed inside such a cathode.

The connecting circuit for a cathode-type valve is given in Fig. 62*c*. Here the heater circuit is quite independent and has no relation to the anode circuit. The cathode is connected to the negative terminal of the anode power supply. In some circuits the cathode is connected to one side of the heater.

In devising circuits with cathode-type valves it should be borne in mind that the insulation between the cathode and the incandescent heater is seldom good enough to withstand a difference of potential greater than 100 volts.

Owing to the large area of the cathode, indirectly heated valves are noted for greater emission in comparison with the filamentary valves. As a rule the qualities of cathode-type valves are superior to those of filamentary valves. In filamentary valves there is no purpose in using thick-diameter filaments, since this would lead to considerably larger filament currents; this is highly undesirable because these valves are generally designed for portable sets, in which the service life of filament batteries is of utmost importance. This is why filamentary valves usually have thin filaments, and hence consume less power than the heaters of cathode-type valves.

The ability of filamentary valves to start operating as soon as they are switched on is also important in portable radio stations. If cathode-type valves are used for such purposes the receiver heaters have to stay switched on even when the transmitter is operating.

This, of course, is intolerable from the viewpoint of the *LT* battery power consumption and explains why filamentary valves, with their immediate "starting up" which is so important in field communication, are generally used in all types of portable equipment. Cathode-type valves are usually employed in medium and high power stationary radio equipment, where it is quite permissible to have the valve heaters switched on all the time because such equipment operates not from batteries but from city mains.

The latest developments in electron valve design have produced new types of cathode possessing high emission capabilities and *high immunity to ion bombardment.* These cathodes are noted for *their comparatively simple* design and high emission current in pulse circuits. Among such cathodes are barium-tungsten cathodes (named the *L*-cathodes), oxide-thorium cathodes, and others.

In all types of valve the cathodes can give a good performance on slight undervoltage, which results in only slightly lowered emission. Such an operating condition is desirable from the viewpoint of increase of valve service life. The only exception, in this case, is the oxide-coated cathodes, which are apt to fail at abnormally low temperatures when they are passing large anode currents. In cathode-type valves in general even slight overvoltage sharply decreases service life without normally improving operation.

30. DIODE CHARACTERISTICS

For all types of valves there is always set a normal and slightly lowered filament or heater voltage, which is maintained at a constant value.

This, however, does not concern the anode voltage, which changes while the valve is in operation. Thus, for instance, in the rectifier circuit of Fig. 61 the anode of the valve is supplied with alternating voltage from the generator, and this voltage is continually changing.

When selecting an electron valve for any type of application, it is important to know the relation between the anode voltage and anode current. *In diode valves this relation is known as diode characteristics,* graphically represented by an appropriate curve.

An example of the characteristic curve is given in Fig. 63*a*. In such a graphic representation anode current I_a is marked off on the ordinate (vertical axis), while the anode voltage U_a is marked off along the abscissa (horizontal axis). Note that I_a is expressed in milliamperes and U_a in volts.

When the anode voltage is zero, the anode current is also zero, because the electrons are not attracted by a neutral anode.

When the anode voltage becomes positive, the anode current begins to increase, at first gradually, then faster. Within certain limits the current increases evenly, then its increase slows down and finally ceases. It will hardly increase any more, no matter how much the anode voltage is increased. This phenomenon is known as *saturation.* The point on the curve where the gradual increase of current stops is known as the saturation point, denoted by I_{sat}.

In Fig. 63*a* the saturation point corresponds to 20 milliamperes, when $U_a = 60$ volts. The effect of saturation is explained as follows. When the anode voltage is low, not all electrons given off by the cathode reach the anode. A considerable part of such electrons

return in the direction of the cathode, surround it, and form *an electron cloud* in the space between the cathode and anode. Such a cloud possesses what is known as *the space charge* (see Fig. 57).

This negative space charge (electrons are negative charges of electricity) repels the new electrons which the cathode tries to emit and interferes with their movement to the anode. If the anode voltage is low, only some of the electrons released by the cathode and travelling at high velocity are able to pierce the space charge and reach the anode. Under such an operating condition the anode current is quite low. This condition, under which the anode is reached only by a part of electrons emitted by the cathode, is called *the limiting condition*, or, in full, the condition of the limiting of the anode current flow by the space charge. Normally all valves operate under the limiting condition. Only at certain times, and then for short periods only, the condition of anode current saturation is reached.

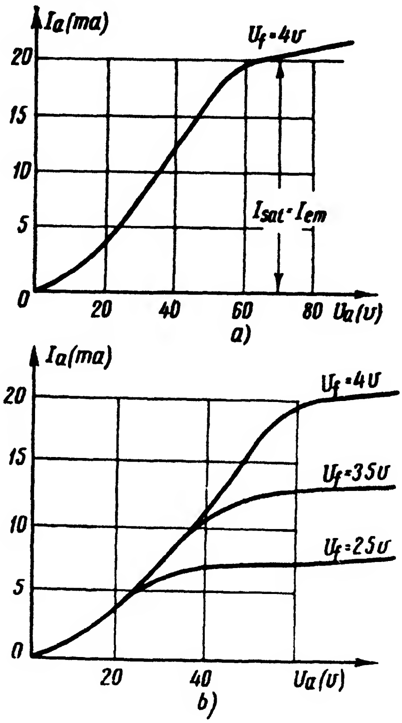

Fig. 63. Diode characteristics

If the anode voltage is made to increase, a constantly increasing number of electrons flying to the anode will be observed, dissipating the electron cloud around the cathode. When the voltage is sufficiently high, all the electrons emitted by the cathode reach the anode, and the electron cloud will disappear. This is the moment when the anode current reaches its saturation value I_{sat}. Obviously the saturation current of the diode is equal to the total emission current I_{em}, the latter being determined by the full number of electrons given off by the cathode every second.

Thus when a diode operates under the condition of saturation, all the electrons emitted by the cathode fly to the anode.

If the temperature of the cathode is raised (filament or heater voltage increased), the emission will correspondingly increase and the saturation current value will be higher. Reducing the cathode temperature decreases the value of saturation current. Fig. 63*b*

gives examples of diode characteristic curves for several values of filament (of heater) voltage U_f.

In modern electron valves the condition of saturation is not a sharply pronounced phenomenon; if U_a is increased after the saturation point is reached, the saturation current does not remain constant but gradually increases. This means that the valve characteristic in the region of saturation is of a rising character. The above effect is attributed to a phenomenon known as autoelectronic emission, when the anode field forcibly tears the electrons from the cathode and when the anode current provides additional heating of the cathode.

The saturation effect is most pronounced in valves with tungsten filaments and least pronounced in those with oxide-coated cathodes, because the electric field of an anode, penetrating into the oxide layer, creates considerable autoelectronic emission. Besides, the oxide layer possesses considerable resistance and is, therefore, particularly prone to additional heating by the anode current.

It should be noted that under an operating condition of $U_a = 0$ the anode current is not totally absent, though in the given case, when there is no positive voltage on the anode to attract electrons, the heated cathode (which emits electrons all the time) gives off its electrons with different velocities, and the fastest of these electrons overcome the repelling action of the electron cloud to reach the neutral anode. The anode current disappears altogether only when the anode is given a slight negative charge (a few fractions of one volt) in respect to the cathode.

In an operating diode the space between the anode and cathode constitutes a peculiar resistor, because it possesses free electrons. Hence we speak of the internal resistance of a valve, usually called the "anode resistance", designated by R_i.

The value of R_i varies, depending upon whether the anode circuit of the valve passes a steady or changing current. The average value of R_i in a diode is several hundred ohms, increasing to thousands and even tens of thousands on low currents corresponding to the lower initial part of diode characteristics.

31. DESIGN AND OPERATION OF TRIODES

A triode is an electron valve with three electrodes. In addition to the anode and emitter, an additional electrode, *a control grid*, is placed in the same valve envelope. This control grid, often called a grid, is made of wire spiral and is located between the cathode and anode. Fig. 64 shows two design versions of triode valves.

The function of the grid is that of controlling the stream of electrons within the valve, i.e., controlling the anode current.

Spacings in the wire spiral allow the electrons to pass freely through the grid, which therefore does not impede the anode current flow. However, in respect to the electric field set up by the positive charge of the anode the grid represents a shield. This field is intercepted by the grid, and only an insignificant part of the field reaches the cathode through the openings in the grid.

Thus the grid shields the cathode from the anode, reducing the influence of the anode upon the electrons emitted by the cathode.

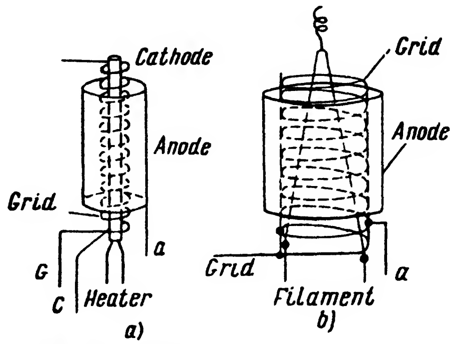

Fig. 64. Design of triode electrodes

Fig. 65 gives a comparison of electrical fields existing in a diode and a triode. As seen from the diagram, the grid intercepts the greater part of the electric field. The finer the grid, the better the cathode is shielded from the influence of the anode. Because of this, and also because the grid is closer to the cathode than the anode, slight changes of grid potential have a much greater effect upon the anode current than that produced by much larger changes of the anode potential.

Grid voltage is the difference of potentials between the grid and cathode. It should be noted that in filamentary valves the grid voltage is measured between the grid and the negative terminal of the filament, the latter being called, as we already know, the common negative.

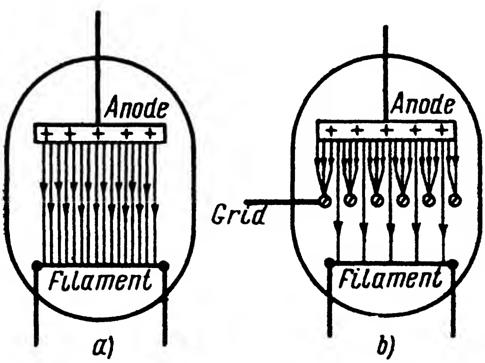

Fig. 65. Electric fields in a diode and a triode

When the grid voltage is slightly negative—i.e., when the grid is negative in respect to the cathode or to the common negative terminal—the grid begins to repel the electrons flowing from the cathode to the anode. However, since the grid voltage is only slightly negative, some electrons manage to fly through it, being attracted by the positive charge of the anode. The electron flow can be stopped altogether only if the grid voltage is made sufficiently negative. In this case the grid will repel all the electrons, thus stopping the anode current. It is then customary to say that the anode current is cut off.

Thus negative grid voltage reduces anode current and can even stop it altogether, creating the condition of cut-off.

The positive grid voltage produces an entirely opposite effect—it helps the anode to attract electrons from the electron cloud. Most of the electrons pass through the grid—although it is also charged

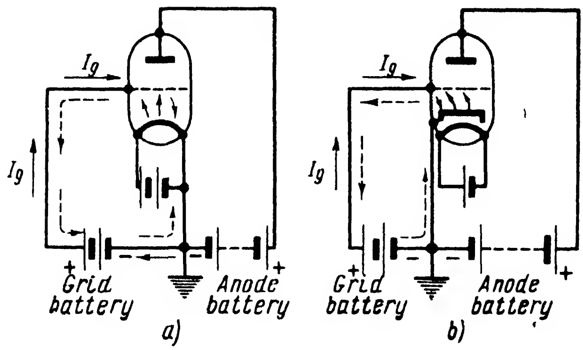

Fig. 66. Circuits of filamentary-type and cathode-type triodes

positively—and, owing to their high speed, rush on to the anode because of their inertia and also because the positive anode voltage is considerably higher than the grid voltage. Still, during this process some electrons are attracted by the grid because of its positive charge and, flowing through the grid circuit, constitute grid current. When the positive grid voltage is sufficiently high, the anode current reaches the saturation point, but the grid current by this time will also reach a high value. Saturation current in a triode is, therefore, smaller than the total emission current of the cathode by a value equal to the value of the grid current.

Thus positive grid voltage increases anode current, can bring it up to the saturation point, and also creates grid current.

Changing the grid voltage from a negative to a certain positive value changes the anode current from zero to the saturation point value. This effect represents the controlling action of the grid.

Fig. 66 gives circuit diagrams of filamentary and cathode-type triodes, including the filament, heater, anode and grid circuits. The grid and all the values which pertain to it are represented by letter g.

In Fig. 66 the positive terminal of grid battery B_g is connected to the grid, and the broken lines indicate the path of grid current I_g electrons. Inside the valve these electrons flow from cathode to grid. In the external wiring of the grid circuit they flow towards the cathode. When grid current is flowing through the cathode wire and through the external circuit of the grid, the total current

$I_c = I_a + I_g$ flows inside the valve and is known as *the cathode current*.

The above-described electron processes taking place in a triode at various grid voltages are illustrated in Fig. 67, where the arrows indicate the movement of electrons.

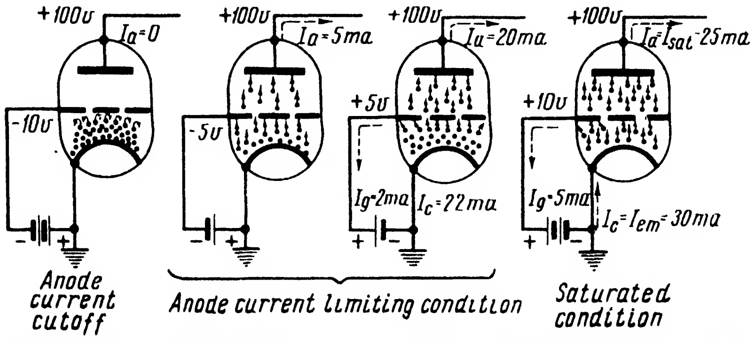

Fig. 67. Movement of electrons in a triode at different values of grid voltage

32. THE TRIODE AS AN AMPLIFIER

Amplification of alternating voltages is one of the main functions of a triode. The alternating voltage to be amplified is applied between the grid and cathode of a triode. The amplified voltage is produced in the anode circuit of the valve. A single-valve amplification circuit is shown in Fig. 68.

Such a circuit, complete in itself, represents a *stage of amplification*. The stage is comprised of a valve, power supplies *HT* and *LT*, and load resistor R_a connected into the anode circuit of the valve.

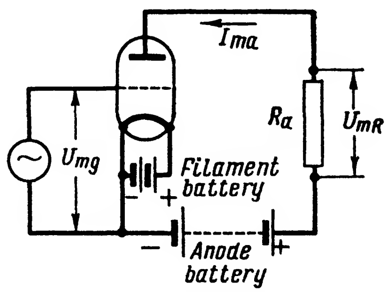

Fig. 68. Amplifier stage circuit

As mentioned above, the alternating voltage signal, fed from some type of device, is applied between the grid and cathode of the valve. This causes pulsations of the valve anode current. The pulsating anode current is comprised of direct-current and alternating-current components. This current, flowing through the load resistor, develops across it a certain pulsating voltage drop, which is also comprised of direct-current and alternating-current components. If the value of load resistor R_a is sufficiently high, the amplitude of alternating voltage U_{mR} across it will considerably

exceed the amplitude of alternating voltage U_{mg} applied across the grid and cathode.

The ratio of these two voltages is called *the amplification factor of a stage* and is denoted by letter k.

$$k = \frac{U_{mR}}{U_{mg}} .$$

The amplification factor of a stage indicates by how much the stage amplifies alternating voltage.

For instance, if $U_{mg} = 2$ volts, and if in the anode current, under the influence of this voltage, has appeared an alternating-current component with an amplitude of $I_{ma} = 0.5$ ma, then if the value of load resistor R_a is 40,000 ohms, the value of U_{mR} will be determined as follows:

$$U_{mR} = I_{ma}R_a = 0.0005 \times 40{,}000 = 20 \text{ volts}.$$

In this case the voltage was amplified 10 times, i.e., the amplification factor of the stage is $\frac{20}{2} = 10$. If, with the same value of I_{ma} ($= 0.5$ ma), R_a were 4,000 (instead of 40,000) ohms, no amplification would take place in the stage, because in such a case $U_{mR} = 0.0005 \times 4{,}000 = 2$ volts. If the value of R_a were still smaller than 4,000 ohms, a decrease of voltage would take place instead of amplification; the output voltage across the load resistor R_a would be smaller than the 2-volt signal voltage applied to the grid circuit.

The above example of calculation gives an approximately correct result because we have not taken into consideration the fact that with the decrease of R_a the alternating current amplitude somewhat increases.

Thus amplification of electrical oscillations takes place in the described amplifier stage; the amplitude of alternating current signal in the anode circuit can be made considerably larger than that of the signal applied to the grid circuit. The energy gain realised in the amplifier stage is made possible not only because of the described action of the valve, but also because of direct current power supplied by the anode battery to the valve circuit. In fact, the energy of electrical oscillations, applied to the grid, controls the conversion of this direct current power into the oscillatory power in the anode circuit of the stage. If the value of voltage applied to the grid is steady, the anode current of the valve has a certain unchanging value. And only when a changing voltage is applied to the grid do pulsations of the anode current begin, the anode current varying in accordance with voltage changes in the grid circuit. It is these anode current pulsations, developing voltage drop across the load resistor, that account for the increase of the oscillatory energy level in the amplifier stage.

The increase of energy of electrical oscillations is the basic property of an electron valve amplifier. In this respect a valve amplifier differs from a step-up transformer, the latter increasing voltage but not the energy level.

A triode amplifier can be used for amplification of low-frequency and high-frequency alternating voltages. Fig. 69*a* gives an example of a high-frequency amplification stage employing an electron valve.

Here the oscillatory circuit L_1C_1, coupled to an aerial and used for receiving radio signals, serves as a generator of alternating voltage, the latter applied across the grid and cathode of an electron valve.

Another oscillatory circuit L_2C_2, tuned to resonance with oscillations being received and amplified, is used as a load resistor in the anode circuit of the valve. On parallel resonance L_2C_2 has a high and purely active (ohmic) resistance.

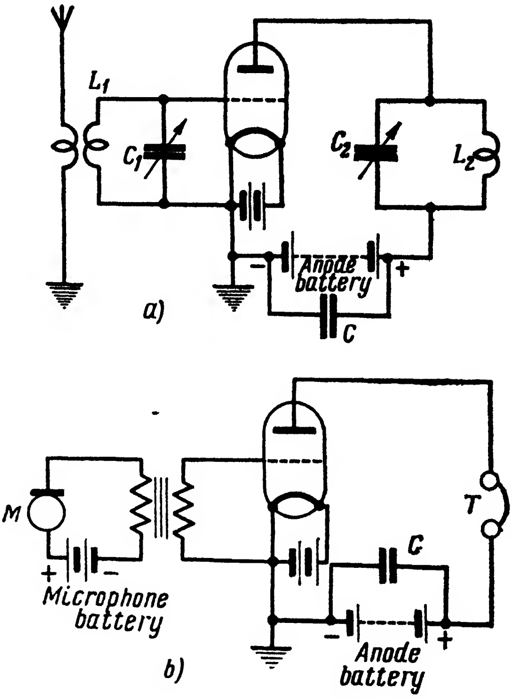

Fig. 69. Circuit diagrams of voltage amplification stages for *(a)* high frequency and *(b)* low frequency

The valve serves as a generator for this tuned circuit. The direct-current component of the anode current freely passes through coil L_2 of the tuned circuit, while the alternating-current component builds up alternating voltage across the tuned circuit, and this voltage will be considerably higher than the voltage built up across the primary (grid) tuned circuit.

The circuit of Fig. 69*a* is an example of the so-called tuned high-frequency amplifiers widely used in radio receivers and transmitters. For reasons which will be explained later, more complex valves are used in such stages instead of triodes.

The circuit shown in Fig. 69*b* is representative of a low-frequency amplifier stage, operating from microphone *M*. Alternating voltage from the secondary winding of the microphone transformer is fed to the grid circuit of the amplifier valve. Earphones *T* are used as

the anode load of the valve. Amplified alternating voltage is built up across those earphones.

Note that by-pass capacitor C is used in both circuits given in Fig. 69. The function of this capacitor is clear from its name; the capacitor, shunting the HT battery and offering negligible reactance to the alternating-current component of the anode current, by-passes this component and does not allow it to flow through the battery.

33. THE TRIODE AS AN OSCILLATOR

As shown in Chapter II, generation of continuous oscillations in a tuned circuit requires periodic addition of electric energy to such a tuned circuit from some external electric power supply.

The periodic application of the external energy to the tuned circuit calls for some special type of extremely quick-acting relay. In this case mechanical relays are quite impractical, but a triode valve can be successfully used in their place. A triode provides the necessary quick action; changes of grid potential offer extremely fast closing and opening of the anode circuit—up to many million times per second, which is far in excess of the operating speeds obtained by the fastest mechanical electromagnetic relays. Up to a frequency of 100 megacycles operation of an electron valve does not experience any retarding inertia effect.

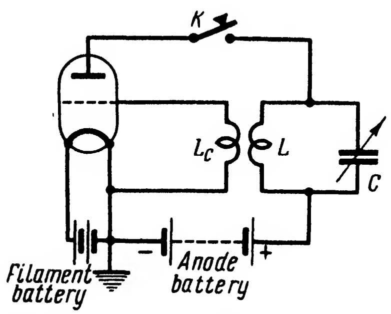

Fig. 70. Circuit diagram of valve oscillator with feedback arrangement

A simple valve oscillator, capable of generating continuous oscillations, consists of a triode, power supplies, tuned circuit LC (Fig. 70) connected into the anode circuit, and coil L_g which is inductively coupled to coil L and connected between the grid and filament of the valve. If the filament of the valve is heated and the anode circuit is closed, the latter will pass current which will charge capacitor C of the tuned circuit. The capacitor will begin to discharge through the coil, setting up free damped oscillations in the tuned circuit. The capacitance and inductance of the tuned circuit will determine the frequency of these oscillations. Alternating current, passing through coil L, sets up alternating voltage in grid coil L_g, the latter voltage controlling the anode current of the valve. When a negative half-wave reaches the grid, the anode current of the valve is cut off. The arrival of a positive half-wave at the grid causes the valve to conduct the anode current. The electrons constituting this

current travel from the anode to the top plate of capacitor C. If the plate is charged negatively at the given moment, the anode current will additionally charge up the capacitor, thus compensating for the losses in the tuned circuit. The described process is repeated every half-period.

However, if the top plate of capacitor C were charged positively during a positive half-wave, the electrons constituting the anode current would reduce the charge on the plate upon reaching it. The oscillations in the tuned circuit would then not be sustained and would be quickly damped out.

To avoid the latter undesirable effect the ends of the coils must be correctly connected. If continuous oscillations are not generated because of wrong coil connection, it is sufficient to merely change over the ends of one of the coils, or to turn one coil through 180° in relation to the other. Besides this, the generation of oscillations requires sufficiently close coupling between coils L and L_g, i. e., the coils must be placed close enough to each other.

A valve oscillator converts the direct current energy of the HT battery into alternating current energy. This process is performed by directing a part of the oscillatory energy from the tuned circuit to the valve grid, thus controlling the anode current, the latter compensating the losses in the tuned circuit. This done, a part of the energy is again transferred from the tuned circuit to the grid circuit of the valve, and energy losses in the tuned circuit are again compensated by the action of the valve and anode battery, etc.

Feedback is the name for the coupling between the anode and grid coils L and L_g, while the process of sustaining continuous oscillations is known as self-excitation.

A common clock is an example of a continuous wave oscillator. Here the pendulum is comparable to a tuned circuit. The wound-up spring acts as a source of energy (anode battery), while the role of relay (valve) is played by the clock mechanism. It is the wound-up clock spring that keeps the pendulum continuously oscillating.

Electron valve oscillators have found very broad application. They are capable of generating currents of almost any frequency, depending upon the L and C values of the tuned circuit. A valve oscillator is an important component of every radio transmitter. Oscillators of this type are also used in many varieties of radio receivers.

34. TRIODE CHARACTERISTICS

In electron valve studies the characteristics of performance curves are of great assistance.

Such characteristics give the dependence of anode current I_a upon grid voltage U_g or upon anode voltage U_a.

*The basic characteristic of a triode is its grid characteristic. This is a curve showing the dependence of anode current upon grid voltage when the anode and filament voltages are maintained at constant values.**

The characteristics of a triode can be plotted with the help of circuit shown in Fig. 71.

In this circuit voltages from the grid and anode batteries are fed to the valve through potentiometers R_1 and R_2, by means of which the voltages U_a and U_g can be varied from zero to the full battery voltages. Voltmeters V_1 and V_2 are used to measure the voltages applied to the grid and anode of the valve. Milliammeter mA reads the anode current of the valve.

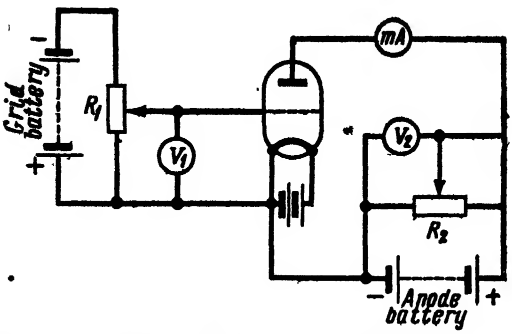

Fig. 71. Circuit diagram for plotting triode characteristics

The plotting procedure begins as follows. Set potentiometer R_2 to a certain position in which a voltage of, say, 100 volts is applied to the anode of the valve. Now operate the grid potentiometer R_1, adjusting it for such a negative value of grid voltage when the valve anode current is cut off. Let the value of this voltage be, for example, —16 v. Next, reduce the grid voltage (moving back the potentiometer slider). This will cause the anode current to appear. Note the value of the anode current for different values of grid voltage.

After taking the anode current reading for $U_g = 0$, change over the polarity of the grid battery and voltmeter V_1 terminals and advance the potentiometer, applying positive voltage to the grid. By recording the values of anode voltage for different settings of potentiometer R_1, the following sort of table will be obtained.

Table 1

U_g (volts)	—16	—12	—8	—4	0	+4	+8	+12	+16	+20	+24
I_a (milliamperes)	0	2.5	7.5	15	22.5	30	37	42.5	45	45	45

On the basis of this table the performance curve shown in Fig. 72 is plotted.

* This characteristic is sometimes called the transfer characteristic.

This curve consists of the following main parts: *lower bend AB, linear middle part BC, upper bend CD, and the saturation region DE.*

The lower bend is attributed to anode current cut off at a certain value of negative grid voltage.

When the grid voltage is adjusted to the value corresponding to the lower-bend operating condition, a particularly dense cloud of electrons is formed near the cathode and strongly opposes the flow of electrons from the cathode to the anode. This is why the anode current changes only slightly in the region of the lower bend.

The upper bend is attributed to the saturation effect which sets in at a certain positive value of the grid voltage (in the given example at +16 volts).

If the grid voltage is made more and more positive, a decrease of anode current can occur. This is known as *an overvoltage operating condition*, which creates the following effect. The total cathode current I_c increases only slightly when the saturation region has been reached, but the positively-charged grid attracts electrons, thus causing a grid current flow.

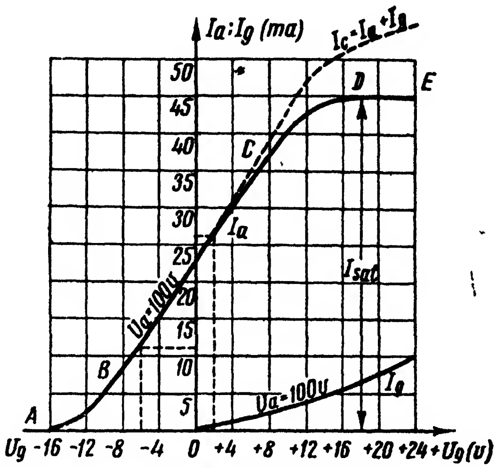

Fig. 72. Grid characteristics of a triode

The higher the positive grid voltage, the greater will be the grid current. This current is part of the total cathode emission and diverts into the grid circuit some of the electrons which would normally flow to the anode of the valve. Hence the decrease of anode current occurs at high positive values of grid potential, when a redistribution of currents takes place.

Anode current is the most important current in an electron valve. In most cases grid current is useless and frequently undesirable.

The performance curve of grid current I_g is given in the same Fig. 72. This curve may be plotted by connecting a milliammeter into the grid circuit. When the grid is made negative, there is no grid current flow. This current appears when the grid potential is zero and increases as the grid is made more and more positive. The broken line in Fig. 72 shows the cathode current I_c curve, this current being equal to the sum of currents I_a and I_g.

Electron valves are seldom operated under conditions corresponding to the upper bend, saturation or grid overvoltage. This is

why operating conditions for these regions are not given in valve catalogues.

Owing to a certain initial velocity of electrons emitted by the cathode of a valve, a small grid current begins to flow when $U_g = 0$. In fact the grid current curve starts when the grid voltage approaches zero but still remains slightly negative (several fractions of one volt).

A plotted performance curve makes it possible to determine the value of anode current for any grid voltage value, when the anode voltage is stipulated. Thus by studying the curve given in Fig. 72 we find that when $U_g = -6$ v, the anode current value will be 11 ma; when $U_g = +2$ v, $I_a = 26$ ma, etc. The dependence of anode current upon grid voltage, when expressed mathematically, is so complex that it is seldom used even by professional radio engineers. This is why graphic representation, such as that given in Fig. 72, is generally used to show such dependence.

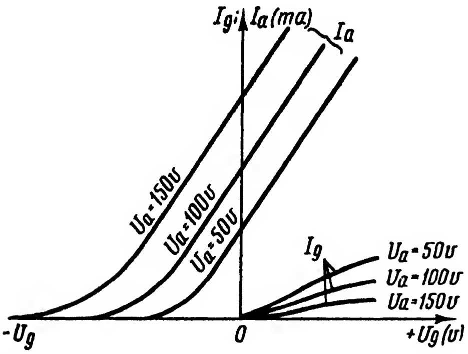

Fig. 73. A family of grid curves of a triode at different anode voltages

The anode current curve given in Fig. 72 is plotted at a certain anode voltage kept constant for all points of the curve. The curve shifts when the anode voltage is changed, although the shape of the curve remains practically the same.

A number of anode current curves may be plotted, each corresponding to a definite anode voltage. Higher anode voltages give greater anode currents for the same values of grid voltage. Accordingly, a curve plotted for higher anode voltage will be located to the left. For lower anode voltages the curve will shift to the right, because the anode current will cut off at lower values of negative grid voltage, and the values of anode currents will be decreased.

Fig. 73 gives a set of performance curves corresponding to different values of anode voltage applied to a valve. Such a set of curves is called *a family of curves*. Apart from anode current characteristics, a family of grid current curves is also given here. It will be noted that the lower the anode voltage, the larger the grid current. When the anode voltage is high, grid current is reduced because a greater number of electrons fly through the grid due to the action of the anode field, grid attraction notwithstanding.

Depending upon the construction of valve electrodes, the grid characteristics of an anode current can be located either to the left

(in the region of negative grid voltages) or to the right (in the region of positive grid voltages). Therefore, valve characteristics (and sometimes the valves themselves) are called *left-handed* or *right-handed*. The finer the grid, the lower is the negative grid potential necessary for the anode current cutoff, and the further to the right are the characteristics shifted. A coarse grid requires higher negative voltage to cut off the anode current, and the characteristics become left-handed. Receiving and amplifying valves are more frequently made with left-handed characteristics, which means that they operate without grid current.

Besides grid characteristics, anode characteristics are often employed to show the dependence of anode current I_a and grid current I_g upon anode voltage U_a, while the grid voltage U_g is kept constant. A family of such characteristics is given in Fig. 74. These characteristics are frequently given only for negative grid voltages, because receiving and amplifying valves generally operate with negative grid potentials to avoid grid currents.

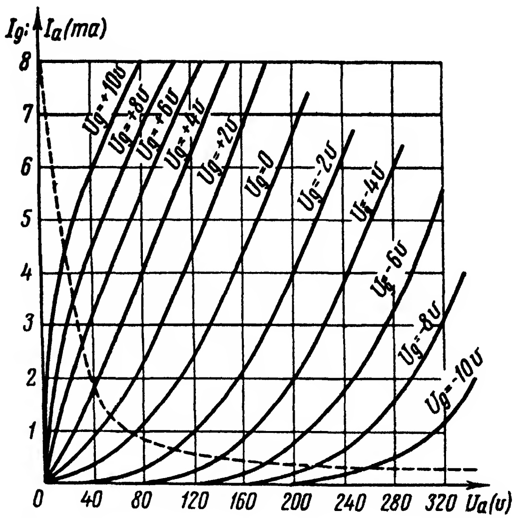

Fig. 74. A family of anode curves of a triode at different grid voltages

The basic anode characteristic for $U_g = 0$ is situated in a similar position to the characteristic of a diode. The curve begins at the point where the anode voltage is zero. At the same point also begin the curves for positive values of grid voltage. These curves are, however, located above the basic characteristic, because a positive grid potential results in greater anode currents. Characteristics for negative values of grid voltage are situated to the right of the basic characteristic and start from respective points corresponding to a definite positive anode voltage. For instance, the curve for $U_g = -4$ v begins at the point corresponding to $U_a = 80$ v. This means that on anode voltages lower than 80 v the anode current is cut off by the negative grid potential —4 v. In a similar way the curve for condition $U_g = -8$ v begins at the point corresponding to $U_a = 160$ v, because the potential of —8 v cuts off the anode current even better.

Anode characteristics also make it possible to determine the values of anode current for different grid and anode voltages. For instance,

performance curves indicate that for conditions when $U_g = -2$ v and $U_a = 120$ v, $I_a = 1$ ma. Anode current I_a increases to 2.2 ma when the anode voltage is raised to 160 v. The broken line in Fig. 74 denotes the grid current characteristic when the grid has a slight positive value.

The curve region of low anode voltages (the region of overvoltage operating condition) is of special interest. In this region the following phenomenon takes place. Electrons, having passed through the grid, are slowed down in the space between grid and anode and in most cases return to the grid. This causes the grid current to increase, while a second electron cloud is formed between the grid and anode. This cloud is dissipated if the anode voltage is increased; the electrons will fly out of the cloud towards the anode. The anode current will then sharply increase, while the grid current drops.

35. TRIODE PARAMETERS

Constant values determining the properties of an electron valve are known as valve perameters. The main parameters of a triode, determining its amplifying abilities, are: *mutual conductance, anode resistance and amplification factor.*

Letter S stands for the value of mutual conductance and shows by how many milliamperes the anode current changes when the grid voltage is changed by 1 volt and the anode voltage remains at a constant value.

Thus the mutual conductance of a valve shows to what extent the grid potential affects the anode current. Mutual conductance is usually expressed in milliamperes per volt (ma/v). If, for instance, a 3-volt change of grid potential brings about a 4.5-milliampere change of anode current (while the anode voltage remains constant), the mutual conductance is found from the following equation:

$$S = \frac{4.5\ \text{ma}}{3\ \text{v}} = 1.5\ \frac{\text{ma}}{\text{v}}.$$

This means that a 1-volt change of grid potential changes the anode current by 1.5 milliamperes. As may be seen from above, mutual conductance is obtained by dividing the value of anode current change by the value of corresponding grid voltage change, the anode voltage remaining steady.

The greater the mutual conductance, the steeper is the grid characteristic. Hence the S parameter is, in effect, representative of the steepness of grid characteristic.

Changes (increment) of a given value are denoted mathematically by symbol Δ. Therefore the following formula may be written to give the mutual conductance:

$$S = \frac{\Delta I_a}{\Delta U_g}, \text{ while } U_a \text{ is constant.}$$

The mutual conductance of a valve is determined by its design. The greater the thermionic emission of the cathode, the closer the grid to the cathode and the finer the grid, the greater will be the S value of the valve. This value varies from 1 to 30 ma/v in valves of different construction.

The S value does not remain the same for different sections of a performance curve; it is maximum on the straight-line section of the curve but is smaller at the upper and lower bends of the characteristic. Typical S values given above and pertaining to different valves hold true for the middle straight-line region of valve characteristics. Valve parameters given in reference charts always pertain to the straight-line region of characteristics lying in the sector of negative grid voltages (for valves with left-hand characteristics).

Mutual conductance can be easily determined from valve characteristics. Fig. 75 gives two characteristics of a valve for anode voltages of 200 and 240 volts. A study of the linear portion of the curve for U_a=200 volts shows that changing U_g from 0 to —2 volts results in a 4-milliampere change of anode current I_a, the latter varying from 9.5 to 5.5 ma. Hence,

$$S = \frac{4 \text{ ma}}{2 \text{ v}} = 2 \frac{\text{ma}}{\text{v}}.$$

For the non-linear part of the curve, the mutual conductance determined in this way is average for the given part. It may be considered that in this case the value of mutual conductance closely approaches the true value in the middle point of this curve section (provided the section is not too steep).

The higher the mutual conductance, the better performance will be given by the valve operating as an amplifier. This is logical, because in such a valve small changes of grid potential will result in large anode current changes, thus producing large voltage drop across the load resistor. Maximum obtainable amplification and also certain other considerations (discussed in the chapter on amplifiers) require that amplifier valves are operated on the linear part of their characteristics, where the value of mutual conductance is greatest.

Anode resistance of a valve, denoted as R_i, is the ratio of anode voltage change to the value of respective anode current change, while the grid voltage is maintained at a constant value.

$$R_i = \frac{\Delta U_a}{\Delta I_a}, \text{ while } U_g \text{ is constant.}$$

In other words, the anode resistance of a valve is the resistance offered to the alternating-current component by the anode-cathode circuit of the valve. The value of anode resistance indicates how the anode voltage influences the anode current, while the grid voltage is steady.

When a triode operates as an amplifier or an oscillator, an alternating anode current component is generated inside the valve as a result of the action of alternating grid potential upon the electron stream. In this case the valve works as an alternating current generator and has a definite internal resistance, just like any other generator.

If a 20-volt change of anode voltage gives a 4-milliampere anode current change (while the grid voltage is maintained at a constant value), the anode resistance of the valve is given by the following: $R_i = \frac{20}{0.004} = 5{,}000$ ohms. If a similar anode voltage change (20 volts) takes place in another valve and causes a 1-milliampere anode current change, R_i of such a valve will be obtained by dividing 20 by 0.001, which gives 20,000 ohms. Thus the less the effect of anode voltage upon anode current, the higher is the anode resistance of the valve.

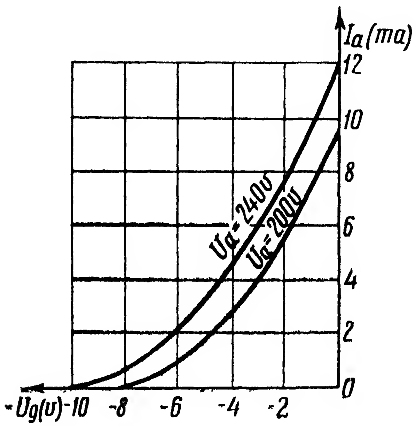

Fig. 75. Grid characteristics of a valve

Various types of triodes have anode resistance values from about 500 ohms (sometimes even less) up to 100,000 ohms. The value of anode resistance is determined by the valve design. The smaller the cathode emission, the finer the grid, the closer the grid is located to the cathode, and the larger the distance between the cathode and anode—the higher will be the anode resistance of the valve.

Different types of circuit call for different values of anode resistance. Thus valves used as high-frequency amplifiers must have high anode resistance if they are not to impair the resonant properties of the tuned circuit connected into the anode circuit of the valve. On the other hand, valves used in low-frequency amplification stages should have low anode resistance.

Let us now see how valve characteristics (Fig. 75) help to determine the value of R_i. Changing the anode voltage by 40 volts (from 200 to 240 v) gives a 2-milliampere anode current change (while the grid voltage is kept constant at —2 volts). In this case the anode resistance is found as follows:

$$R_i = \frac{40}{0.002} = 20{,}000 \text{ ohms.}$$

The value of R_i is approximately the same for all points of the linear part of the performance curve and is at minimum, increasing in the upper and lower bends.

Anode resistance R_i of a valve should not be confused with direct-current resistance R_0 of the given valve. The latter value is not a constant parameter and changes even over the linear part of the valve characteristic. The R_0 value is also determined by Ohm's law; in this case we divide anode voltage U_a by anode current I_a. For instance, referring to the characteristics given in Fig. 75, we find that the anode current is equal to 6 milliamperes when $U_a=240$ volts and $U_g=-3$ volts; hence $R_0=240:0.006 = 40{,}000$ ohms. Taking the same anode voltage of 240 v but a different value of grid operating condition, where $U_g=0$, we find that the anode current is equal to 12 milliamperes; here the value of R_0 is respectively decreased and becomes equal to: $R_0 = 240:0.012=20{,}000$ ohms.

The amplification factor, denoted by Greek letter μ, shows how much greater is the effect of grid voltage change upon the value of anode current than that of anode voltage change upon the same current.

For instance, if a 40-volt anode voltage change gives a 1-milliampere anode current change, and if the same anode current change is obtained by changing the grid potential by 2 volts, it becomes clear that the grid voltage effect upon the current is 20 times as large as the anode voltage effect and, hence, the amplification factor of the valve is equal to 20.

The grid of an electron valve has a more pronounced effect upon the anode current, chiefly because the grid stops the larger part of the electric field set up by the positive charge of the anode and hence weakens the action of the anode upon the current. When the anode is positive in respect to the cathode, electrons emitted by the cathode are attracted by the anode because of the electric field set up between the anode and cathode. The grid acts as an obstacle (shield) to this field. The finer the grid, the greater will be the portion of the electric field stopped by it, and the weaker will be the attraction of the electrons by the anode. At the same time, the electric charge of the grid itself will act upon the electrons given off by the cathode without any weakening because there is no obstacle to the electric field between the grid and cathode.

The strength of the electric field can be conventionally considered in terms of the number of lines of force. If the grid stops 90% of electric lines of force set up by the anode and passes the remaining 10% of such lines to the cathode, it may be said that the grid weakens the action of the anode 10 times. In this case the anode acts 10 times more weakly than the grid, and the amplification factor of the valve is equal to 10. A finer grid will stop, say, 95% of the lines of force and will pass only 5% (one-twentieth part). The amplification factor of such a valve will be 20, because the action of its anode is 20 times weaker than the action of its grid. The value of the amplification factor of a valve is chiefly determined by the construction of its grid; the finer the grid, the greater is the amplification factor μ.

Amplification factor is the value determining alternating voltage amplification by an electron valve. As an example let us consider an amplifier employing a valve with amplification factor of 10. Assume that application of alternating voltage with an amplitude of 2 volts to the grid of the valve results in obtaining an alternating-current component of 2 milliamperes in the anode circuit of the same valve. In other words, a 2-v voltage change in the control grid circuit produces a 2-ma change in the anode current. If an external alternator with an output voltage of 2 v is connected directly into the anode circuit of the valve, the alternating-current component amplitude will be 10 times smaller (0.2 ma). To obtain, in this case, an alternating-current component of 2 ma in the anode circuit, we would have to employ an alternator with an output voltage of 20 volts, instead of 2 volts. But, as we have already seen, the required alternating anode current of 2 ma was obtained when a 2-volt alternating signal was fed to the grid of the valve. Consequently, applying alternating voltage to the grid produces a varying current in the anode circuit, and the value of such current is the same as that of the current obtained from an external alternator connected into the anode circuit of the valve, if the voltage of the alternator is ten times higher than the alternating voltage applied to the grid. This example illustrates that application of a 2-volt alternating signal to the grid of a valve is equivalent to connecting a 20-volt alternator into the anode circuit of the valve. Thus the valve itself acts as an alternator, connected into the anode circuit and developing, in the given case, an e.m.f. of 20 volts.

It may be considered that feeding alternating voltage U_{mg} to the grid of a valve makes the valve operate as an alternator, the output voltage of which is μ times higher than U_{mg}. Such a valve, acting as a generator of alternating anode current, is fed with energy from the anode power supply. Such a conception of an electron valve as an alternator was first offered by M. A. Bonch-Bruyevich and later, independently of him, by the German scientist G. Barkhausen.

So the amplification factor μ of an electron valve really shows by how many times the valve can amplify alternating voltage. Owing to the internal resistance of a valve, it is, however, impossible to utilise in full the alternating e.m.f. obtained in the anode circuit. In order to secure the amplified voltage, a high load resistance R_a is connected in series with the valve anode. A part of the alternating e.m.f. generated by the valve is lost in this load resistor. Another part is always lost in the anode resistance of the valve itself. The ratio of amplified alternating voltage U_{mR}, obtained across the load resistor R_a, to alternating voltage U_{mg}, applied to the control grid of the valve, is the amplification factor of the whole amplifier stage. This factor is denoted by letter k. At the same time the ratio of the full alternating e.m.f. developed in the anode circuit to the alternating voltage applied to the grid is the amplification factor

of the valve itself. Hence the amplification factor of a valve is always greater than the amplification factor of the whole stage. If the valve had no internal (anode) resistance, the whole of the alternating e.m.f. would be fully utilised in the load resistor, and the amplification factor of the stage would be equal to that of the valve. However, such an ideal situation is never observed in practice.

The following numerical example will confirm the above statements. Assume that a certain valve has the following parameters: $\mu = 10$; $R_i = 10{,}000$ ohms and is used in conjunction with a 40,000-ohm load resistor R_a. If a signal voltage $U_{mg} = 2$ v is applied to the grid, an e.m.f. of $2 \times 10 = 20$ v will act in the anode circuit. This e.m.f. will be distributed between R_i and R_a in proportion to their values. Alternating voltage appearing across R_a will be equal to $U_{mR} = 16$ v, i.e., to 80% of the whole e.m.f. (because 40,000 ohms is 80% of the full resistance of the anode circuit, which is 50,000 ohms). In this case the amplification factor k of the whole stage is given by $k = 16:2 = 8$, which is less than μ of the valve (10). Of course, the greater R_a in comparison with R_i, the larger the part of the alternating e.m.f. of the anode circuit that will be acting across R_a and the closer to μ will be the value of k.

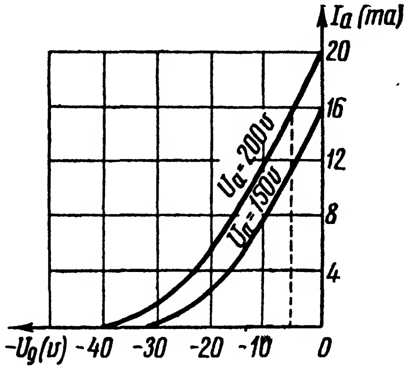

Fig. 76. Grid characteristics of a valve

Let us now refer to valve characteristics for the purpose of determining amplification factor. Fig. 76 gives valve characteristics for anode voltages of 150 to 200 v. As seen from the curves, when grid voltage is maintained at a constant value of, say, 0 volts, a decrease of anode voltage by 50 volts (from 200 to 150 v) is accompanied by a 4-ma anode current decrease, the current value dropping from 20 to 16 milliamperes. If the anode voltage is kept constant and equal to 200 volts, a similar 4-ma reduction of the anode current can be obtained by changing the grid voltage from 0 to -5 v. Thus a 50-volt anode voltage change gives the same effect, as far as the anode current is concerned, as a 5-volt grid voltage change. Apparently the action of the grid is 10 times stronger than the action of the anode. The amplification factor is equal to 10 and is represented by the ratio of anode and grid voltages giving a similar anode current change ($\mu = 50:5 = 10$).

The value of μ is determined by the following formula:

$$\mu = \frac{\Delta U_a}{\Delta U_g},$$

where ΔU_a and ΔU_g are, respectively, anode and grid voltage changes affecting the grid current to a similar degree.

Depending upon their design, different triodes possess amplification factors whose values vary between 4 and 100.

Sometimes instead of amplification factor μ its reciprocal $\frac{1}{\mu}$ is used. This reciprocal is denoted by letter D and shows what part of the electric field generated by anode voltage penetrates through the grid to the cathode. If, for instance, $\mu = 10$, then $D = \frac{1}{10}$, which indicates that only one-tenth of the electric field set up by the anode charge passes through the grid.

There is a very simple relation between the basic parameters S, R_i and μ. This relation, called the internal equation of a valve or Barkhausen's formula, makes it possible to determine any one of the above three parameters when the other two are known. The relation is expressed as follows:

$$\mu = R_i S.$$

In this formula the mutual conductance must be given in amperes per volt. It is also possible to express S in milliamperes per volt, providing that R_i is given in kilo-ohms. This formula is easily checked by substituting values of ΔU_a, ΔU_g and ΔI_a for its parameters and by dividing the right-hand side by ΔI_a.

Example 1. Find the amplification factor of a valve with the following parameters: $R_i = 20{,}000$ ohms; $S = 4$ milliamperes per volt.

Solution:

$$\mu = 20{,}000 \times 0.004 = 80.$$

Example 2. A valve has the following parameters: $\mu = 25$; $S = 2$ milliamperes per volt. Find its anode resistance.

Solution:

It follows from formula $\mu = R_i S$ that $R_i = \mu : S = 25 : 0.002 = 12{,}500$ ohms.

As shown above, the parameters of a triode can be determined from its grid characteristics. Such parameters can be also found from the anode characteristics.

Besides the three basic parameters, each type of valve is also characterised by values of normal supply voltages, emission current, service life, maximum permissible anode dissipation, and other factors.

Maximum permissible anode dissipation, designated by $P_{a\,max}$, depends upon the design, dimensions and material of the anode. This is a very important parameter for any type of valve. The value of anode dissipation is expressed in several fractions of one watt for small receiving valves, but it runs into many kilowatts in the case of high-power transmitting valves.

In an electron valve the electrons constituting the anode current move at high velocities owing to the attraction of the anode, striking the latter with considerable force. The higher the anode voltage, the higher becomes the velocity of the electrons. For instance, when the anode voltage is equal to 100 volts, the electrons are travelling with velocities as high as 6,000 kilometres per second at the moment they strike the anode. As a result of such electron bombardment the anode is heated to red or white heat and may even melt. The power dissipated by the anode (P_a) can be determined by the following formula:

$$P_a = I_a U_a.$$

For instance, if $U_a=30$ volts and $I_a=20$ milliamperes, the anode of the given valve will dissipate 0.6 watt.

This power is considered as lost because heating of a valve anode produces no useful effect. Excessive anode heating is highly undesirable, for the anode can melt or else liberate gases, disturbing the vacuum.

Excessive heating of the valve anode can be avoided if the following condition is observed: $P_a<P_{a\ max}$; i.e., when the power actually dissipated by the anode is smaller than the maximum permissible power. $P_{a\ max}$ is increased by enlarging the surface and general dimensions of anodes in electron valves, by making the anodes of high-heat metals and by supplying the anodes with special cooling ribs. Besides these measures, anode surfaces are blackened, because black surfaces give off heat better than light-coloured ones. High-power valves employ running-water cooling, first suggested by Bonch-Bruyevich, or else are cooled by forced-air ventilation.

36. DYNAMIC OPERATING CONDITION OF VALVES

When a load resistor is connected into the anode circuit of an electron valve, the operating condition of such a valve is said to be *dynamic*. When no such load resistor is provided in the circuit, it is customary to say that the valve operates under a *static* condition. The grid and anode characteristics of a triode and its basic parameters S, R_i and μ, studied above, pertain to the static operating condition. They should therefore be referred to as static characteristics and static parameters. For simplicity sake the word "static" is usually omitted.

In radio equipment valves are usually operated under a dynamic condition, with load resistors connected into the anode circuits. Only in cases when the resistance values of these load resistors are small in comparison with the anode resistance of the valves with which they are used would it be correct to consider the valves operating under an approximately static condition (e.g., we have such a condition when only a milliammeter or another device with similarly

negligible resistance is connected in the anode circuit of an operating valve).

The dynamic operating condition of a valve is noted for the varying anode voltage when alternating voltage is applied to the grid. This is different from the static operating condition, where the anode voltage U_a is equal to the anode power supply voltage E_a and remains constant during fluctuations of the grid potential, although the latter causes anode current changes. Under a dynamic operating condition the anode voltage of a valve is smaller than the voltage of the anode power supply, because a part of the latter voltage is dropped in the load resistor,

$$U_a = E_a - I_a R_a.$$

When an alternating voltage is applied to the grid of a valve, the anode current begins to pulsate, the voltage drop across the load resistor varies and, hence, the anode voltage U_a changes, too. During this process anode voltage changes are always in phase opposition to grid voltage changes.*

Let us assume that the grid voltage has increased, i.e., became more positive in respect to cathode potential. This will cause an increase of the anode current and an increase of voltage drop across the load resistor, while the anode voltage will be correspondingly decreased—i.e., changed in the opposite direction as compared to the grid voltage change. Thus anode voltage changes counteract grid voltage changes. When a grid voltage change causes an increase of anode current, the lowering of anode voltage accounts for a certain decrease of the same current. When a grid voltage change causes a decrease of anode current, the action of the anode voltage counteracts this, too, to a certain extent. This is why under the condition of dynamic operation the anode current changes are less pronounced than under static operation, because in the latter case there is no effect of anode counteraction, sometimes called anode reaction. Under the condition of dynamic operation, the mutual conductance of a valve is reduced as compared to that during static operation.

The dynamic characteristic is of considerable assistance in the study and calculations of valve operation under dynamic conditions and can be easily plotted when a family of static characteristics is available and when the values of anode supply voltage E_a and load resistance are given. Fig. 77*a* gives an example of plotting a grid dynamic characteristic of a certain valve for conditions when $E_a = 200$ v and $R_a = 5{,}000$ ohms.

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* We are studying the dynamic operating condition of a valve only for the case when the load resistor is represented by pure ohmic resistance. If the load also possesses some reactance, additional phase shifts will occur. Such a case is more complex and is only seldom experienced in practice.

If $U_g = -12$ v, the anode current of the valve is cut off. Therefore the starting point A of the dynamic characteristic coincides with the initial point of the static characteristic when $U_a = 200$ v. At lower values of negative grid voltage anode current again appears and produces voltage drop across load resistor R_a, and this decreases the anode voltage. Available static characteristics are employed for plotting the dynamic characteristic in the following way. At a

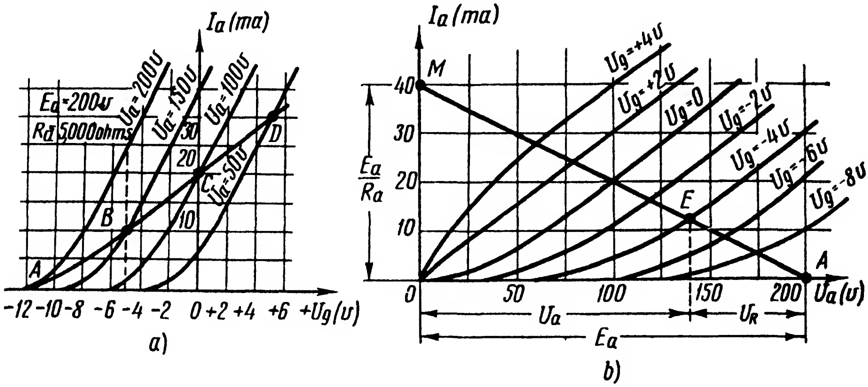

Fig. 77. Plotting grid *(a)* and anode *(b)* dynamic characteristics of a triode

certain value of anode current anode voltage $U_a = 150$ v; i.e., the voltage drop across R_a will be equal to 200—150 = 50 v. In this case the anode current may be found by Ohm's law:

$$I_a = \frac{E_a - U_a}{R_a} = \frac{50}{5{,}000} = 0.01 \text{ a} = 10 \text{ ma}.$$

Point B corresponds to this current value. The transition to this point is accomplished by decreasing the negative grid voltage from —12 v to —5 v.

In a similar manner point C is found for $U_a = 100$ v. This point corresponds to the following current value:

$$I_a = \frac{100}{5{,}000} = 0.02 \text{ a} = 20 \text{ ma}.$$

For $U_a = 50$ v, point D is obtained and corresponds to the following current value:

$$I_a = \frac{150}{5{,}000} = 0.03 \text{ a} = 30 \text{ ma}.$$

Connecting all these points with a smooth curve, we obtain a grid dynamic characteristic showing the changes of anode current and anode voltage in relation to the grid voltage, together with the difference between static and dynamic operating conditions. On static operation U_a is always equal to E_a and to 200 v; therefore

changing the grid voltage from —12 to —5 v produces anode current change of 25 ma. On the other hand, on dynamic operation the current increases only by 10 ma (from point A to point B), because during the process the anode voltage is reduced by 50 v. As may be seen the steepness of the dynamic characteristic is much less than that of the static characteristic. The greater the value of load resistor R_a, the larger will be the anode voltage change and the smaller will be the anode current change, i.e., the dynamic mutual conductance becomes smaller and the performance curve passes lower.

When studying the performance of various valves working under a dynamic operating condition we shall make use of the dynamic characteristic, representing it without the family of static curves.

A dynamic characteristic can be also plotted among a family of anode static curves. In this case the plotting procedure is simpler because the anode dynamic characteristic is a straight line. Let us assume that a family of anode static characteristics is available (Fig. 77*b*) and it is required to plot a dynamic characteristic for known values of E_a and R_a. For the sake of comparison Fig. 77*b* shows the anode characteristics of the valve whose grid characteristics were given in Fig. 77a. For consistency we also retain the former values $E_a = 200$ v and $R_a = 5$ kilo-ohms.

The plotting procedure is carried out with the help of equation $U_a = E_a — I_a R_a$. This is an equation of the first power in respect to U_a and I_a and, hence, in the system of co-ordinates the curve is represented by a straight line, which stands for the anode dynamic characteristic (sometimes also called the load line). Such a characteristic can be plotted with the help of two points and it is convenient to use those at which the line intersects the X-axis and Y-axis. If in the equation given above we take $I_a = 0$, then $U_a = E_a$ (point A).

In this case the anode current of the valve is cut off by a considerable negative grid voltage. Point A corresponds to the initial point A of the grid dynamic characteristic (Fig. 77*a*). The second point is determined from the same equation, if U_a is taken equal to zero. Then, solving the equation for I_a, we find that $I_a = \frac{E_a}{R_a}$. In our example $I_a = \frac{200}{5} = 40$ ma. Marking off this value along the Y-axis, we obtain point M. The required dynamic characteristic is the straight line drawn through points A and M.

Unlike point A, which corresponds to the real operating condition of the valve (cutoff), point M does not correspond to any actual condition of operation, because when $U_a = 0$ the anode current cannot have a maximum value but must be equal to zero. Hence point M is required only for the plotting of the dynamic characteristic, but this point, as well as the AM section of the line near this point, do not correspond to the actual dynamic characteristic

(apparently point M corresponds to a condition when the anode and cathode of the valve are short-circuited to each other).

The anode dynamic characteristic serves to illustrate a certain effect, i.e., how a change of grid voltage into positive direction causes an increase of anode current, accompanied by a decrease of anode voltage because of increased voltage drop across the load resistor. When such a characteristic is available, it becomes possible to find anode current, anode voltage and voltage drop across the load resistor U_R for any value of grid voltage.

Thus if $U_g = -4$ v, the operating condition of the valve is determined by point D. Referring to the curve we obtain the following for this point: $I_a = 12$ ma, $U_a = 140$ v, and $U_R = 60$ v. When I_a is equal to 12 ma, the other values obtained from the curve must be really as indicated, because: $U_R = I_a R_a = 12 \times 5 = 60$ v, while $U_a = E_a - U_R = 200 - 60 = 140$ v.

Under conditions of static operation, when there is no load resistance R_a, the anode voltage is always 200 in the given example. The current will be 30 ma when $U_g = -4$v. In dynamic operation current decrease down to 12 ma is caused by the lowering of anode voltage from 200 to 140 v.

All other conditions being equal, the greater the resistance value of R_a, the smaller will be the current through the valve and the lower will be situated the dynamic characteristic (point M will be located lower). However, point A, standing for the cutoff, will remain in the same place because it is determined solely by the voltage of power supply (E_a).

Since anode dynamic characteristics are plotted very simply and offer convenient means of determining the values of U_a and U_R, they are frequently employed in various electron valve calculations.

37. RECEIVING AND AMPLIFYING TRIODES

The triodes described below are used in radio receivers and amplifiers. Those used in more powerful equipment (oscillators and transmitters) will be dealt with later in Chapter VIII.

All the valves carry abbreviated designations, consisting of letters and figures. The previous system of marking, still used for certain glass valves, has the following significance:

The first letter—field of valve application (У — amplification; П — radio signal reception; Т — amplification in sound diffusion equipment; С—special application, entailing special construction of the valve).

The second letter — type of cathode (filament) (О — oxide-coated, Б — bariated).

After these two letters comes a figure denoting the production assignment number.

Example: СБ-240 valve, i.e., a special-application valve provided with bariated cathode and manufactured as a part of a lot with the production assignment number 240.

The new system of marking uses different symbols; when referring to receiving and amplifying valves, first comes a figure showing a rounded-off value of filament or heater voltage; then a letter indicating type and application of the valve; letter C indicates that the valve is a triode. Then comes a figure which is different for valves whose other letters and figures are similar but which have different characteristics and parameters; and finally a letter indicating the type of valve design. Typical designations for various designs are:

С — standard-size glass valve;

П — miniature ("bantam") glass valve;

Ж — special-type miniature ("acorn") valve designed for ultra-high frequency applications;

Л — "loctal" valve with a special locking device, firmly holding the valve base to its socket;

Б — miniature valve provided with a 10-mm diameter envelope;

А — miniature valve provided with a 6-mm diameter envelope;

Д — valve designed for ultra-short waves and provided with special disc-type or cylindrical terminals.

When the last letter is omitted this indicates a metal valve. For instance, designation 6C5 indicates a metal triode designed for heater voltage of 6.3 volts; 6C1Ж stands for an "acorn" glass triode with a filament voltage of 6.3 volts.

Old glass valves were provided with four-pin bases (Fig. 78*a*), in which the pin terminals were located unsymmetrically. The farthest pin was connected to the valve anode and was placed opposite to the control-grid pin. If the valve was of the cathode type, its cathode was connected to an additional pin located in the centre of the base (Fig. 78*b*).

Modern glass and metal valves employ an eight-pin "octal" base (Fig. 79). The name "octal" indicates the location of base pins in a

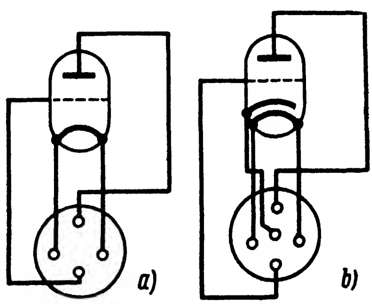

Fig. 78. Base connection of old-type glass triodes

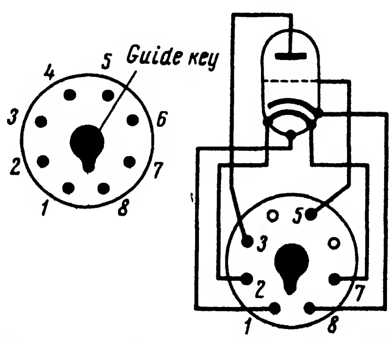

Fig. 79. Location of pins in an octal valve base, valve 6C5 serving as an example

regular octagon. In the centre of such an octal base is installed a longer guide pin (key) of considerable thickness. The guide key is made of insulating material and secures correct insertion into the socket.

The pins of an octal valve are numbered as shown in Fig. 79. In a number of types of octal valve the heater is connected to pins *2* and *7*, while the cathode to pin *8*. Pin *1* is connected to the body of metal valves or, in metallised glass valves, to the metal layer covering the glass envelope. The anode of such a valve is frequently connected to pin *3*.

It should be particularly noted, however, that in many types of metal valves the electrodes are connected to the pins in a different order from that shown above. Hence it is best to refer to the valve base connection diagrams given in valve catalogues and references whenever dealing with any type of valve.

Only few types of valves actually need all the eight pins offered by the standard octal valve base. Hence in many valves unnecessary pins are not provided, though the octal socket retains its usual shape and dimensions.

Fig. 79 shows base connections of type 6C5 metal triode. This valve uses an octal base but is provided only with those pins which are shown blackened in the drawing.

Although valve manufacturing plants produce triodes of filamentary and cathode-type varieties, metal valves are all of the latter variety.

In accordance with their application, triodes are classified as follows: (a) low-power high-μ triodes; (b) low-power triodes with medium μ; (c) high-power triodes with low μ.

Let us now examine the general aspects of modern electron valve design. First note that the dimensions of valve envelopes are determined by the power which the valve has been designed to handle. In higher-power valves the electrodes are of considerable dimensions and are capable of radiating a great amount of heat. This heats the valve envelope, and it becomes necessary to increase its surface.

In glass valves the electrodes are supported on a mount, shaped as a glass stem flattened at one end. Platinite conductors are sealed in this mount and are used for connecting valve electrodes to the terminal pins (platinite is a nickel-iron alloy with the same thermal expansion coefficient as the glass). The upper ends of the platinite conductors are welded to thicker conductors which carry the actual electrodes (Fig. 80*a*). The lower ends of the platinite conductors are jointed to copper wires which run to the contacting pins of the valve base.

A glass tube, passing through the mount, is used for pumping out air from the valve envelope and is sealed after the operation. In some types of valves the air is pumped out through a "pip" provided right in the glass envelope. The thick electrode-carrying

conductors are fixed on mica plates above and below the electrodes; this helps to keep the electrodes at a definite distance from each other. These mica plates are firmly fixed in the glass envelope (Fig. 80*b*) and, securing a constant positioning of the electrodes, assure high-degree of constancy of valve parameters.

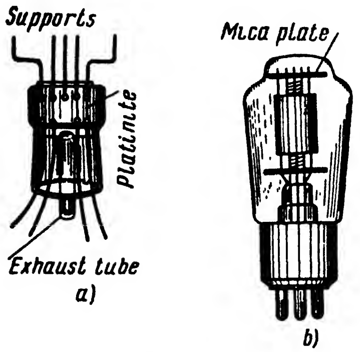

Fig. 80. Electrode-supporting mount and external view of a glass valve

Metal valves have the following advantages over glass valves: smaller dimensions, greater mechanical strength, higher constancy and uniformity of parameters, and good shielding of electrodes from various external electric and magnetic fields, such shielding being provided by the valve envelope itself.

Strong heating and greater possibilities of air penetration into the envelope are the disadvantages of metal valves; therefore only certain low-power valves are now of the metal variety.

The envelope of a metal valve is made of steel. In previous metal-valve manufacturing practice a metal disc was welded to the lower part of the envelope. Such discs were made of "fernico" (nickel-iron alloy resembling platinite), provided with holes sealed with glass drops (Fig. 81*a*). Fernico conductors, passing through the glass drops, served as the electrode supports. An evacuating pipe, made of metal, was placed in the guiding key.

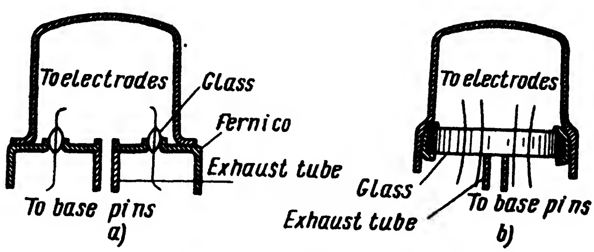

Fig. 81. Electrode terminals of a metal valve

Modern valve-manufacturing practice prefers to employ a solid glass bottom, sealed in a fernico ring welded to the metal envelope of the valve, with terminal wires passing directly through the glass bottom (Fig. 81*b*). If the valve is provided with a top terminal, the latter is also passed through a glass insert sealed in a hole made in the upper part of the envelope.

Special "acorn" valves are produced for operation on frequencies of 300-600 megacycles (Fig. 82*a*). The electrodes of such valves have very small dimensions, because it is very important on such frequencies to keep the inter-electrode capacitances to minimum possible values. The spaces between the electrodes of acorn valves are also made as small as possible in order to decrease the transit time of electrons flying from one electrode to another. Such transit time must be less than an oscillatory period, which is very small on extremely high frequencies. The electrode terminals in such valves are brought out directly through the glass envelope and represent short and straight pieces of wire.

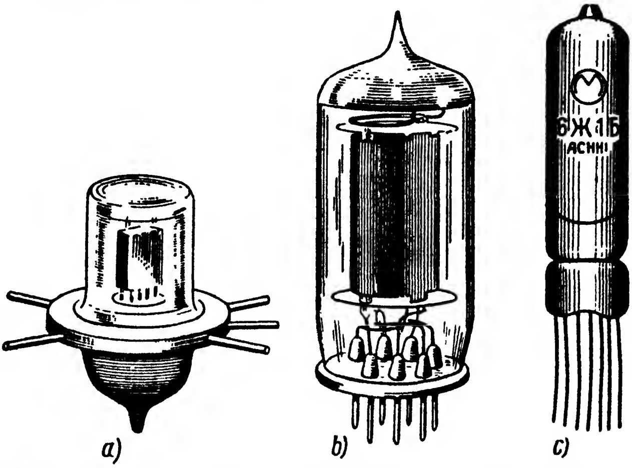

Fig. 82. External view of valve types: *(a)* acorn, *(b)* bantam, *(c)* baseless miniature

"Bantam" valves (Fig. 82*b*) is the name for miniature glass valves with no bases and intended for operation on frequencies up to 200 mc. In valves of this type as many as seven and even nine terminal wires, connecting valve electrodes to external circuits, are passed right through the thick glass bottom. Bantam valves are more convenient in installation than acorn valves. They are at present produced in filamentary and cathode-type versions, designed for 1.2 v and 6.3 v operation respectively.

The most popular types of valves are: low-power triodes 6Ф5, 6C5, and 6C2C, acorn triode 6C1Ж, and bantam triodes 6C1П and 6C2П. Among the 2-v miniature valves manufactured up to low-power triodes УБ-240 and 2Ф2М should be noted. Power output triodes used in receivers and amplifiers are: УО-186 (4-v series), 6C4C and 2C4C.

Recently the production of new cathode-type miniature and baseless valves has begun. These (Fig. 82*c*) are designed to operate on 6.3 heater voltage (for instance 6C6Б triode and others). In the new

miniature valves 0.4-mm tinned wires are used as electrode terminals, brought out of the glass envelope side by side. The envelopes of such valves carry tags giving the type of valve and location of various terminals, and at one end of the glass mount through which the wire terminals pass in a row a coloured mark is provided to enable counting off the leads in correct succession.

When such valves are employed in high-frequency stages of ultra-short-wave equipment, the electrode terminals are cut as short as possible to secure their connection to external circuits.

38. DISADVANTAGES OF TRIODES

Triodes are noted for two major disadvantages.

The first is insufficient amplification. This arises from an insufficient degree of shielding between anode and cathode, because the control grid allows a significant part of the electric field of the anode to penetrate through it. If a grid with very finely-spaced spiral is installed in a triode, the electrons emitted by the cathode will encounter great difficulty in penetrating, and consequently the anode current will be extremely small and will be easily cut off by comparatively small values of grid voltage. Practically the whole characteristic of such a valve will be located in the region of positive values of grid voltage, when the grid current is extremely large, which is highly undesirable. A high-μ triode always has a right-hand characteristic. In practice the μ value of a triode does not exceed 100.

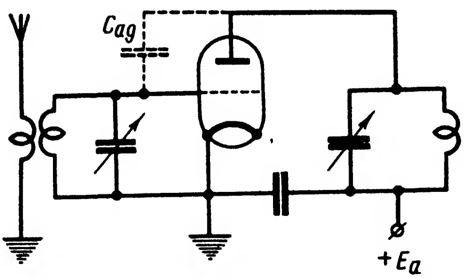

Fig. 83. Anode-grid capacitance in a high-frequency amplifier stage

Considerable capacitance existing between the anode and grid of a triode is the second disadvantage. Capacitance always exists between the electrodes of a valve as between any metallic conductors separated by a dielectric (as, e.g., by a vacuum). In an electron valve the value of such capacitance may be only several picofarads, but this is enough to have considerable effect on the valve operation.

Capacitance between the anode and control grid of an electron valve, usually referred to as *transfer capacitance*, produces particularly harmful effects in high-frequency amplifier stages (Fig. 83). It causes capacitive feedback, which is highly undesirable in an amplifier. Sometimes called parasitic capacitance, it is capable of making an amplifier stage behave like a self-excited oscillator. If a high-frequency amplifier begins to generate its own oscillations

("parasitic oscillations"), its proper operation is upset because its only duty should be that of amplifying signals delivered to it from another source. Besides, if such self-oscillating high-frequency amplifier is a part of a radio receiver, its oscillations are radiated by the receiver aerial, creating interference to other receivers, which is also very undesirable.

Therefore a triode cannot be used as a high-frequency amplifier in radio receivers. However, triodes may be employed in low-frequency amplifier stages, where several picofarads of parasitic capacitance can cause no great harm on audio frequencies because such low capacitance offers a high value of capacitive reactance, causing only a very small value of feedback.

It should be noted that two other inter-electrode capacitances also exist in a triode, and these capacitances also sometimes lead to undesirable effects. One such capacitance exists between the grid and cathode and is known as the *input capacitance.* It uselessly loads the source of voltage being amplified and connected between the grid and cathode of the amplifier valve.

On low frequencies the impedance of the input capacitance is very high and has no noticeable effect upon the operation of a valve. On higher frequencies, however, the capacitive reactance of the input circuit decreases, allowing greater values of alternating current to flow through the input capacitance. This leads to a considerable voltage drop in the internal resistance of the source whose voltage is to be amplified, and consequently reduces the useful alternating voltage applied to the grid of the amplifier valve. The higher the frequency and the higher the internal resistance of the voltage source, the more pronounced is this harmful phenomenon.

Output capacitance is that existing between the anode and cathode of an electron valve. Connected in parallel with the load resistor of the valve, it reduces the total resistance of the load and decreases amplification of the stage. As in input capacitance, the shunting effect of the output capacitance is hardly noticeable on low frequencies but begins to tell when the frequency increases and capacitive reactance of the output circuit is reduced.

The effects of the input and output capacitances need not be taken into consideration in tuned amplifiers where oscillatory circuits are provided both at the input and output ends of the stage, which is generally the case in high-frequency amplifiers and oscillators (see Figs 69*a* and 70). Here the input and output capacitances simply cause a slight increase in the total capacitance of both tuned circuits. Everything said above about the input and output capacitances of a triode also holds true for more complex valves, the study of which we shall now begin.

89. DESIGN AND OPERATION OF A TETRODE

The disadvantages of a triode can be considerably lessened by the insertion in the valve of a fourth electrode shaped as a grid and placed between the anode and control grid.

This additional electrode is called a screen grid, and such a four-electrode valve is known as a tetrode.

In a tetrode the screen and control grids shield the cathode from the electric field of the anode. The amplification factor μ of the valve is thus considerably increased, while the parasitic capacitance between the anode and control grid becomes quite insignificant.

Let us explain this phenomenon by the following example. Assume that the amplification factor of a triode is equal to 20 and the parasitic anode-grid capacitance of the same valve equals 10 picofarads. Now let us insert into the valve a screen grid, whose openings pass only one-fiftieth part of the electric field set up by the anode. Thus the number of lines of force reaching the control grid will be reduced to one-fiftieth of the previous value. The control grid itself will pass only one-twentieth part of such lines.

Consequently only one-twentieth part of one-fiftieth of the field will reach the cathode. The influence exerted by the anode will be accordingly weakened to 0.001 of its previous value, and the amplification factor will be 1,000.

Thus the insertion of a screen grid into the valve has increased the amplification factor 50 times and, at the same time, decreased the parasitic anode-control grid capacitance by the same amount, the latter dropping to a mere 0.2 of a picofarad.

To obtain a similar amplification factor ($\mu = 1{,}000$) in an ordinary triode we should have to use such a fine control grid which would pass only one-thousandth part of lines of force. However, such a fine grid would pass practically no anode current, and the whole design would be useless. On the other hand neither of the two grids in a tetrode are too fine, and the electrons pass through both of them with comparative ease (the screen grid is actually made finer than the control grid).

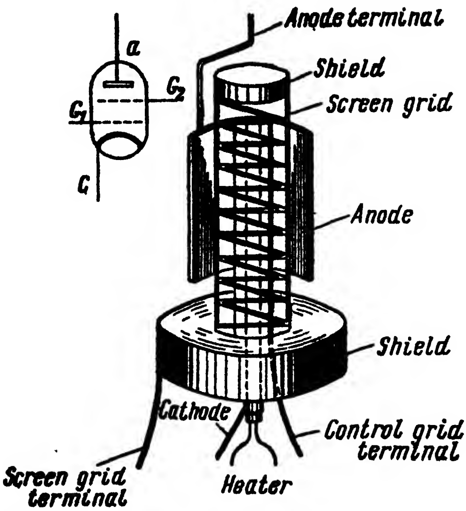

Fig. 84. Electrode construction of a tetrode and its graphic representation in circuit diagrams

Some of the lines of force given off by the anode may still reach the control grid, passing not through the screen grid but in a round-about way; this would result in an undesirable increase of the parasitic capacitance between the anode and control grid. To avoid this metal screens are also installed in tetrodes to intercept such lines of force, in addition to the installation of screen grids. Fig. 84 gives a schematic representation of a tetrode and shows the construction of tetrode electrodes (for better clarity a part of the anode is cut away).

Harmful parasitic capacitance may also be created between the terminals of the anode and control grid, as well as between the anode and control grid wiring. To avoid this, anode and control grid terminals in high-frequency tetrodes are separated from each other as far as possible, or else additional shields are provided between them. The anode and grid wires, external to the valve, are also shielded from each other. Shielding methods consist of running control grid wires through the metal sheathing, providing special screens between anode and grid leads, covering glass valves with shielding metal covers, and thoroughly earthing various shielding devices to the chassis of the radio set, which in turn is connected to external earth and to the common negative terminal of the electric circuit.

40. TETRODE CONNECTIONS

The screen grid of a tetrode must not be connected to the valve cathode (nor to the common negative terminal). This is easily understood if we consider that there must be a difference of potential between the screen grid and cathode if we want the emitted electrons to move towards the anode. The anode itself attracts the electrons weakly because its field has to act through two grids. Hence a direct connection between the screen grid and cathode will result in very low anode current.

In order to make a tetrode operate properly, a positive voltage U_{g2} must be applied to the screen grid. This voltage is called screen-grid voltage, and its value is usually between 20-50% of the anode voltage value. Under the influence of screen-grid voltage, screen-grid current I_{g2} will flow, being comprised of electrons attracted by the positively-charged screen grid.

There are three general methods of supplying positive voltage to the screen grid of a valve. The simplest is applying voltage from a part of the anode battery to the screen grid (Fig. 85*a*). However, it may happen that the anode battery has not the required taps, and this method is seldom used. A more popular means is connecting the anode battery to the screen grid through a dropping resistor R_{g2} (Fig. 85*b*), the value of such resistor reaching several tens of

thousands or even hundreds of thousands of ohms (in case of low-power tetrodes). Screen-grid current I_{g2}, flowing through such resistor, produces a voltage drop across it, this drop being subtracted from the anode battery voltage E_a:

$$U_{g2} = E_a - I_{g2}R_{g2}.$$

If the screen-grid current corresponding to the given operating condition of a valve is known (it can be found from valve characteristics or tables), then the value of dropping resistor R_{g2} can be calculated with the help of the following formula:

$$R_{g2} = \frac{E_a - U_{g2}}{I_{g2}}.$$

Thus, if $E = 160$ v, $U_{g2} = 60$ v and $I_{g2} = 0.5$ ma, then

$$R_{g2} = \frac{160 - 60}{0.0005} = \frac{100}{0.0005} = 200{,}000 \text{ ohms}.$$

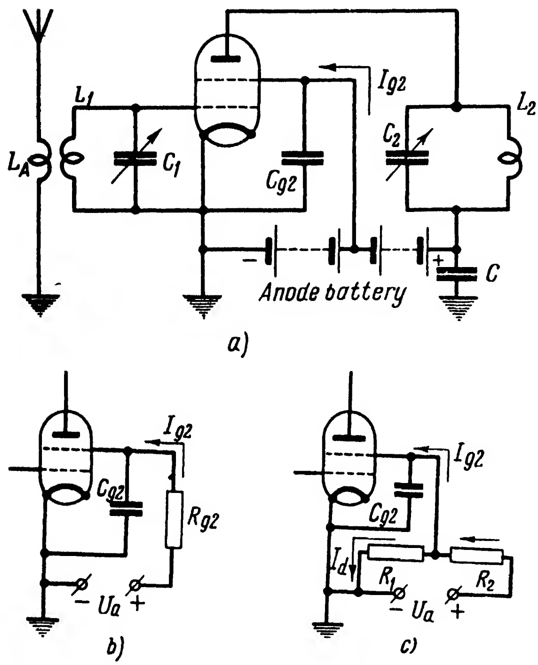

Fig. 85. Methods of supplying positive voltage to the screen grid of a tetrode

Another quite popular method of supplying positive voltage to the screen grid of a valve consists in using a voltage divider (Fig. 85*c*). In this case the anode battery is connected across a voltage divider comprised of two series-connected resistors R_1 and R_2 with a total value of several tens or thousands of ohms, which pass the divider current I_d. The voltage created by this current across resistor R_1 is fed to the screen grid.

This method is less economical than the second described above because a part of the anode battery energy (current I_d) is wasted. Voltage dividers are therefore seldom used in battery-type radio receivers.

Despite its poor economy a voltage divider arrangement has its own advantage. When screen-grid voltage is supplied through a series dropping resistor (as in the second method), various changes of current I_{g2}, caused, for example, by filament or anode voltage changes, result in sharp changes of screen-grid voltage U_{g2}. Such changes are considerably less when a voltage divider is employed,

since in such an arrangement the distribution of voltage across resistors R_1 and R_2 depends not only upon current I_{g2} of the screen grid, but also upon voltage-divider current I_d, which is almost independent of the operating condition of the valve. The greater the value of current I_d in comparison with current I_{g2}, the more stable will be screen-grid voltage U_{g2}, but the higher will be the waste of anode battery power spent on heating the divider. Usually the value of current I_d is adjusted to several milliamperes.

Calculation of resistance values R_1 and R_2 can be performed on the basis of Ohm's law, providing the values of E_a, U_{g2}, I_d and I_{g2} are known.

$$R_1 = \frac{U_{g2}}{I_d} \text{ and } R_2 = \frac{E_a - U_{g2}}{I_d + I_{g2}}$$

Example: Calculate a voltage divider which will deliver $U_{g2} = 40$ v to the screen grid of a valve from anode battery with voltage $E_a = 120$ v, if screen-grid current $I_{g2} = 1$ ma and divider current $I_d = 3$ ma.

Solution:

$$R_1 = \frac{40}{0.003} = 13{,}300 \text{ ohms;}$$

$$R_2 = \frac{120 - 40}{0.003 + 0.001} = \frac{80}{0.004} = 20{,}000 \text{ ohms.}$$

If a screen grid is to eliminate parasitic capacitance between anode and control grid, it must be connected to the common negative terminal (cathode) through a capacitor of sufficiently large capacitance. Such a capacitor must possess a very low capacitive reactance and must act as a short-circuit on alternating current. The capacitance of a capacitor used for this purpose should be between several thousands or several tens of thousands of picofarads when used in a high-frequency stage. This capacitor C_{g2} is shown in all circuits using tetrodes (Fig. 85). If it is not employed in a tetrode circuit, the alternating current is likely to pass from the anode circuit of the tetrode to its control grid circuit, flowing through two series-connected inter-electrode capacitances, the anode-screen grid capacitance C_{ag2} and screen grid — control grid capacitance C_{g2g1} (Fig. 86). In this case the parasitic feedback between the anode and control grid circuits will not be eliminated.

When capacitor C_{g2} is provided in the circuit, high-frequency alternating current from the tuned anode circuit will pass through anode-screen grid capacitance C_{ag2}, the low-reactance capacitor C_{g2}, and the capacitor C back to tuned circuit L_2C_2, thus avoiding the control grid circuit.

Capacitor C_{g2} also performs the following additional function. When a tetrode is employed in an amplifying circuit, the screen-grid current pulsates like the anode current, and, if the alternating-current component of the screen grid passes through resistance R_{g2}, the voltage drop across this resistance will also be of pulsating nature and will cause instability of the screen-grid voltage. Capac-

itor C_{g2} by-passes the alternating-current component of screen-grid current, flowing directly to the cathode of the valve; this component therefore does not flow through resistance R_{g2}, and consequently voltage U_{g2} remains constant, which is important if the tetrode is to operate properly.

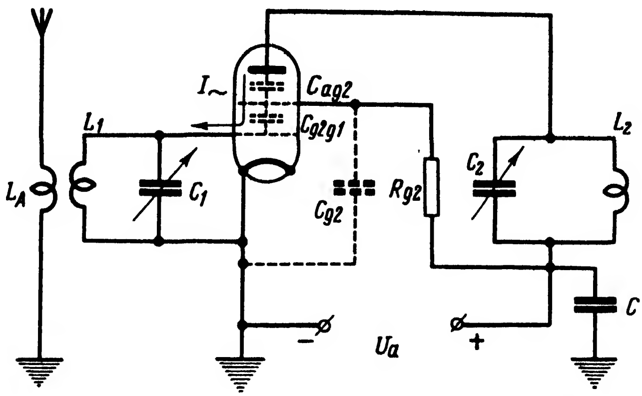

Fig. 86. Parasitic feedback in a tetrode when the screen-grid capacitor is omitted

41. GRID CHARACTERISTICS AND PARAMETERS OF A TETRODE

Fig. 87 gives grid characteristics of a tetrode for various anode and screen-grid voltages. Each pair of anode current characteristics, located very close to each other, corresponds to a definite screen-grid voltage. When the anode voltage varies from 200 to 150 volts, the characteristic is shifted only slightly because, owing to the shielding effect of the two grids, the anode exerts only a very slight influence upon the anode current.

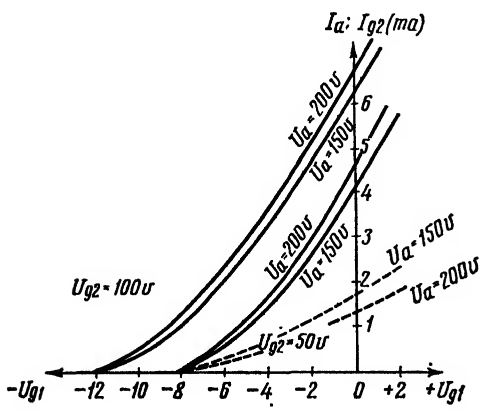

Fig. 87. Grid characteristics of a tetrode (or a pentode)

If, however, a similar voltage change (50 volts) takes place in the screen-grid circuit, a sharp shift of characteristics will take place because the action of the screen grid is weakened only by a comparatively coarse control grid.

The broken lines in the drawing represent screen-grid current characteristics for $U_{g2} = 50$ v and for anode voltages of 150 and 200 volts. As may be seen, on lower anode voltage the screen-grid current increases because the screen grid attracts a greater quantity of electrons. Anode current and screen-grid current characteristics start at the same point, i.e., both are cut off simultaneously. Really, under the condition of cutoff, there is no anode nor screen-grid current. And if some electrons pass through the control grid, a part of them will inevitably reach the screen grid, while the remainder will fly through the screen-grid openings and be attracted by the anode.

Despite its high amplification factor, the characteristics of a tetrode are "left-handed", because the positive voltage of the screen grid acts through a comparatively coarse screen grid and, in order to secure the condition of cutoff, a considerable negative potential has to be applied to the control grid.

For various anode voltages, grid characteristics of a tetrode are a diverging bunch of curves. This is attributed to the following. Anode voltage changes only slightly affect the total (cathode) current, because the anode field acts through two grids. However, such anode voltage changes bring about a redistribution of the total electron stream. For instance, when U_a is made to increase, the screen-grid current is reduced while the anode current is correspondingly increased. Let us assume that raising the value of U_a from 150 to 200 volts results in a 10% increase of the anode current. If the anode current has a value of 1 ma, such an increase will constitute 0.1 ma, while for 10-ma value of anode current the increase would amount to 1 ma. Simultaneously with this action, screen-grid current is decreased by the same amount.

This accounts for the fact that the divergence between characteristics increases when the value of anode current is raised. Because the characteristics corresponding to various values of anode voltage pass close to each other, the dynamic characteristic only slightly differs from static characteristics, and its steepness is only a little more pronounced than that of static curves.

Tetrode parameters are noted for the following peculiarities. Mutual conductance is within the same limits as in the case of a triode (from 1 to 30 ma/v), but the amplification factor of a tetrode reaches several hundred units. The anode resistance of a tetrode is also higher than that of a triode and reaches several hundred thousand ohms.

Tetrode characteristics are more curved than those of triodes. Because of this, and also because of the diverging character of the curves, tetrode parameters are not so constant as those of triodes.

42. DYNATRON EFFECT IN TETRODES

The phenomenon of secondary emission in tetrodes is a serious disadvantage of this type of valve. Electrons emitted by the cathode and striking the anode knock out *secondary electrons* from the latter. It is even possible that each striking electron can dislodge more than one secondary electron. This is called *secondary emission* and is observed in all valves. In diodes and triodes the effect is not noticeable because the secondary electrons have a low velocity, do not fly to a considerable distance from the anode, and return to it under the influence of its positive charge.

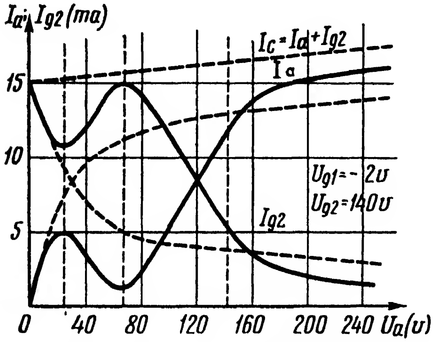

Fig. 88. Anode characteristics of a tetrode

In a tetrode secondary emission is also not objectionable — provided that the screen-grid voltage is lower than the anode voltage, for under such conditions all the secondary electrons return to the anode. However, when a tetrode operates with large signal amplitudes, its anode voltage goes through large variations because a voltage increase across the anode load resistance (due to the action of anode current) causes a decrease of anode voltage. Thus there may occur certain moments in the operation of a tetrode when its anode voltage drops below the constant screen-grid voltage. It is during such moments that the secondary electrons dislodged from the anode will not return to it but will be attracted by the screen grid because of its greater positive potential.

A stream of secondary electrons is thus created, flowing in the opposite direction to the anode current (the latter constituted of primary electrons). This is known as the *dynatron effect*, owing to which the total anode current of the valve is reduced.

The expression "current of secondary emission" should not be confused with the "effect of secondary emission". The presence of secondary emission is required for the effect to set in, but this dynatron effect actually occurs only when the screen-grid voltage value begins to exceed the value of the anode potential.

The setting in of the dynatron effect can be easily observed in the anode characteristic shown in Fig. 88 and pertaining specifically to a tetrode. As the anode voltage is increased, the anode current at first also increases because at low velocities the primary electrons do not dislodge secondary electrons from the anode. Then begins the secondary emission and the anode current decreases. With further

increase of anode voltage the current of secondary emission decreases, while the anode current begins to increase again. When the anode voltage value begins to exceed the screen-grid voltage, the effect of secondary emission does not cease, but it is no longer evident because the secondary electrons, dislodged from the anode, return to the anode and no longer fly to the screen grid. In this case it may be observed how secondary electrons, dislodged from the screen grid, are attracted to the anode. This causes an additional increase of the anode current, while the screen-grid current is somewhat decreased.

This creates a "dip" in the anode current characteristic; a "drooping" section is formed in the curve and within the limits of this section an increase of anode voltage results not in raising but in the lowering of the anode current.

Varying the screen-grid current produces the opposite effect. Lowering the anode current corresponds to increasing I_{g2}, and vice versa. The broken lines in Fig. 88 show the behaviour of characteristics in the absence of secondary emission. The same drawing also gives the characteristic of total (cathode) current which, when the anode voltage is raised, increases only slightly because of slight influence of the anode.

In amplifiers the dynatron effect upsets the proper operation of valves and makes it impossible to amplify large-amplitude signals.*

Suppression of the dynatron effect requires that screen-grid voltage in tetrodes be always considerably lower than the anode voltage. This, incidentally, is also desirable from the viewpoint of power economy; screen-grid current, doing no useful work and being a waste of electrical power, should be kept as low as possible by operating the screen grid at minimum reasonable voltage values.

43. DESIGN AND OPERATION OF A PENTODE

The harmful dynatron effect in a tetrode may be prevented by inserting between the anode and screen grid of the valve yet another grid called the *suppressor grid*. *Five-electrode valves including a suppressor grid among their electrodes are called pentodes*. Pentodes, because of their advantages, have found a wide application in various radio equipment.

The suppressor grid is connected to the cathode (filament) of a valve and is therefore of the same potential as the emitter. Like the cathode, it is charged negatively in relation to the anode and, consequently, repels secondary electrons dislodged from the anode,

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* Despite this there are also useful applications of the dynatron effect (see Section 84).

opposing their movement in the direction of the screen grid even when the positive potential of the latter exceeds the anode potential. Introduction of the suppressor grid eliminates the harmful dynatron effect.

The suppressor grid has quite a coarse spiral but nevertheless provides additional weakening of the anode field. In many types of pentode the suppressor grid is connected to the cathode inside the. valve, and no external connection is provided for it. When a

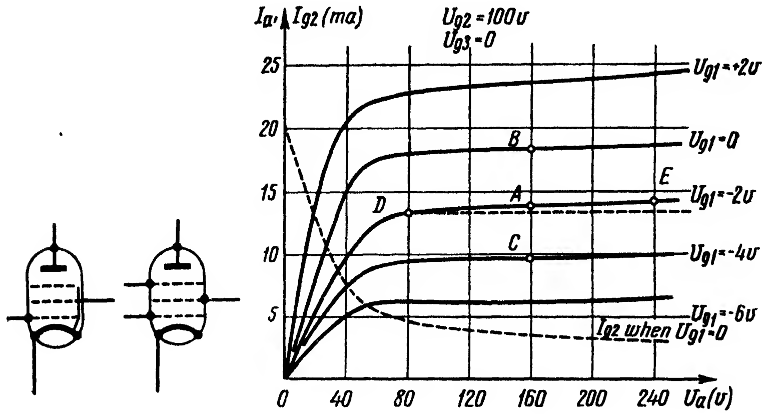


Fig. 89. Schematic representation of a pentode and its anode characteristics

valve has a separate suppressor grid terminal, external connection between it and the cathode is made right at the socket when the circuit is being wired (Fig. 89).

As far as valve parameters are concerned, pentodes differ from tetrodes in that they have a still higher amplification factor μ, which reaches several thousand units in certain types of pentode. This is explained by the additional anode field shielding which the suppressor grid provides (the third obstacle to the anode field). Hence the control grid provides an even stronger action upon the anode current in a pentode than it does in a tetrode.

Owing to the additional shielding provided by the suppressor grid, parasitic anode-control grid capacitance C_{ag1} is also lower in a pentode than in a tetrode. The mutual conductance of pentodes is about the same as that of triodes and tetrodes. The anode resistance of pentodes is quite high, reaching hundreds of thousands and, in some cases, millions of ohms.

High-frequency pentodes employ screen grids with a finer spiral and consequently have a high amplification factor (several thousands), high anode resistance (up to several megohms), and low anode-grid capacitance. The screen grid of a low-frequency

pentode is somewhat coarser; hence the amplification factor and anode resistance are lower than those of high-frequency pentodes, and their anode-grid capacitance is not decreased quite so much.

The grid characteristics of a pentode have about the same shape as those of a tetrode (see Fig. 87); at different values of anode voltage pentode characteristics are still closer to each other, because the anode action in a pentode is weaker than in a tetrode.

Fig. 89 gives anode characteristics of a pentode for various control grid voltages at a constant screen-grid voltage. At first the characteristics rise steeply, i.e., small changes of anode voltage produce a sharp increase of anode current.

This is explained as follows. When the anode voltage is zero the electrons, under the influence of high screen-grid potential, fly through the spaces of the screen grid, whereupon their velocity is slowed down because the screen grid begins to attract them back to it. This makes them stop and return to the screen grid. An accumulation of electrons — the second electron cloud — is formed between the screen grid and the suppressor grid.

The anode acts upon the electrons of this cloud through the coarse suppressor grid; therefore a slight increase of anode voltage gives a sharp rise of anode current. If U_a continues to increase, the electron cloud is dissipated and the anode current increase is slowed down. Finally, all the electrons that passed through the screen grid are attracted to the anode, which has a sufficiently high potential. As a result, the electron cloud disappears. The anode current rise experienced upon further anode voltage increase is attributed mainly to the increase of the number of electrons attracted by the anode from the electron cloud hovering in the vicinity of the cathode. In this case the anode acts through the three grids and its action is weakened hundreds and even thousands of times. This is why a considerable anode voltage increase causes only a very small change of anode current. The characteristics lose their steepness and become almost horizontal, and these parts are used in setting the operating condition of the valve, because on them a pentode has a high amplification factor and high anode resistance, while in the initial steep parts of the characteristic the amplification factor, mutual conductance and anode resistance have comparatively low values (about the same as in triodes).

The higher the negative voltage of the control grid, the lower the anode current and the lower the characteristic. The peculiarity of the anode characteristics of a pentode is that they will be less steep and closer to each other as the negative control grid voltage increases. This corresponds to the increase of anode resistance and to the decrease of mutual conductance at higher negative value of control grid voltage. The broken line in Fig. 89 shows the screen-grid current characteristic for one value of control grid voltage.

The studied characteristics offer a proof of the absence of dynatron effect in a pentode.

The level part of an anode current characteristic in a pentode (or a tetrode) should not be confused with the saturated condition. In these types of valve, due to the great weakening of the anode action by the three grids, it is practically impossible to reach the saturation current by means of anode voltage increase. It can be obtained in this case only at high positive values of control grid voltage.

The analysed family of anode characteristics corresponds to a definite voltage of the screen grid. The anode and screen-grid current characteristics are shifted higher as the screen-grid voltage increases. However, if the latter voltage considerably exceeds the anode voltage, a redistribution of currents will take place; there will be a sharp increase of the screen-grid current, while the anode current will decrease.

The determination of pentode parameters from its characteristics has certain peculiarities. The grid characteristics are located so near each other that they make it possible to determine only the mutual conductance of the valve. Accordingly the parameters of a pentode are found from anode characteristics given in reference tables. Let us find, for example, the parameters for point A, corresponding to $U_a = 160$ v and $U_{g1} = -2$ v (Fig. 89). In this case the mutual conductance is determined by finding the increment of the anode current and grid voltage between points B and C. Here we find that $\Delta I_a \approx 8$ ma and $\Delta U_{g1} = 4$ v. Hence, $S \approx 8:4 = 2$ ma/v. The anode resistance is determined from the increment of the anode voltage and anode current between points D and E. These values are, accordingly, $\Delta U_a = 160$ v and $\Delta I_a = 1$ ma, and then $R_i = 160 \div 1 =$ $= 160$ kilo-ohms. The amplification factor is found from the formula $\mu = SR_i$ and is equal to $2 \times 160 = 320$.

The connection of a pentode in the circuit of an amplifier is in principle the same as that of a tetrode. The suppressor grid is connected to the cathode inside the valve or externally. Methods of feeding the screen grid are the same as in a tetrode. In low-power amplifiers screen-grid voltage is made equal to 20-50% of the anode voltage because the anode current is rather small in amplification of weak signals. When strong signals are amplified, the anode current must be large, and in this case the screen-grid voltage can be equal to the anode voltage and can even exceed it for higher-power pentodes. Pentodes designed for oscillator duty operate on screen-grid voltages which are equal to 20-80% of anode voltage. In cases when the screen-grid voltage must be equal to anode voltage, this grid is connected directly to the positive terminal of anode power supply, using no voltage divider or dropping resistor. In low-frequency amplifiers capacitor C_{g2} must have a considerable capacitance (0.1 mfd and higher); this makes its capacitive reactance sufficiently low to low-

frequency currents. All that was said above about the shielding of a tetrode also pertains to pentodes working as high-frequency amplifiers.

As stated previously, the suppressor grid voltage of a pentode is usually equal to zero. However, in pentodes operating as oscillators a small value of positive or negative voltage is frequently applied to suppressor grids. In such circuits the suppressor grid functions in the usual manner because its voltage is always considerably lower than the anode voltage of the valve. Application of positive voltage to the suppressor grid of a pentode operating as an oscillator helps to increase the useful power of the stage. Negative voltage is fed to the suppressor grid of a pentode when this grid is used for modulating high-frequency oscillations generated by an oscillator.

Sometimes a pentode is employed as a triode. In such cases the screen grid is connected to the anode terminal, these two electrodes operating as one. If the suppressor grid has a separate terminal, it is also connected to the anode. On triode connection of a pentode, characteristics and parameters of the valve are accordingly altered; the anode resistance and amplification factor of the valve are lowered, but the mutual conductance remains approximately the same.

Pentodes are the most universal and widely used type of valve. They have found application in almost every kind of radio apparatus.

44. BEAM TETRODES

Besides the pentodes, the so-called *beam tetrodes* have also received a wide recognition. In a beam tetrode the dynatron effect is eliminated not with the help of a suppressor grid but in another way.

Beam tetrodes are noted for the following peculiarities of design. The distance between the screen grid and anode is made large, and the control grid and screen grid have the same number of turns, placed opposite each other (lined up), as shown in Fig. 90*a*. Owing to such alignment of screen grid and control grid turns, the electrons emitted by the cathode travel in sheets to the anode (Fig. 90). To prevent them from travelling in the direction of the holders supporting grid turns, special screens S_1 and S_2 are provided (Fig. 90*b*). These screens are connected to the cathode and are at zero potential. Therefore the electrons can travel only in the directions shown in the drawings.*

A schematic representation of a beam tetrode is given in Fig. 90*c*, but this type of valve is frequently represented as an ordinary tetrode.

* As an additional measure, in modern beam tetrodes parts of the cathode directed towards the grid holders are not covered with oxide-coat layer.

When the screen-grid voltage is greater than the anode voltage in a beam tetrode, the primary electrons, being "braked down" in the space between the anode and the screen grid, move slower. This leads to the formation of a negative charge in a certain part of this space, such a space charge "braking down" the secondary electrons flying out of the anode and returning them to the anode.

Thus the space charge formed by the electrons themselves plays the role of a suppressor grid.

In ordinary tetrodes the electron stream is dispersed by the holders and turns of the screen grid, which is made finer than the control grid; therefore the electrons cannot travel to the anode in condensed sheets or "beams", as a result of which there is no sufficiently dense charge, forming in the space between the anode and the screen grid, capable of repelling the secondary electrons back to the anode. In this respect beam tetrodes have an advantage over ordinary tetrodes. Their other advantage is the low value of screen-grid current they consume, explained by the fact that the electrons travel chiefly through the spacings of the screen grid and are not stopped by it.

Beam tetrodes are used in output stages of low-frequency amplifiers and in radio transmitters. The characteristics of beam tetrodes are quite similar to those of pentodes. Fig. 90*d* shows the family of anode characteristics of a beam tetrode. In comparison with those of a pentode they are noted for sharper transition from the

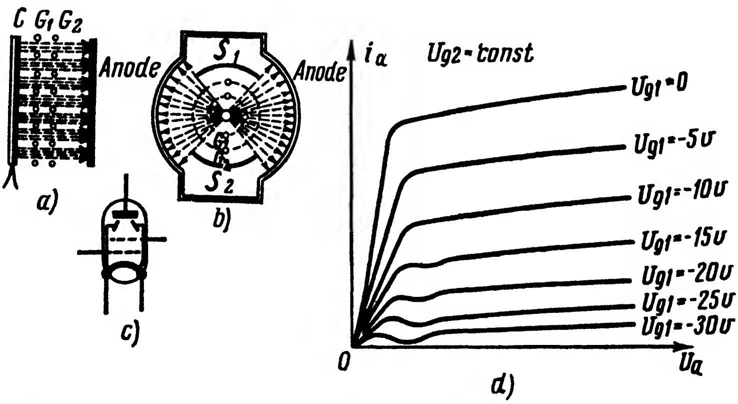


Fig. 90. Construction of electrodes of a beam tetrode, its schematic representation and anode characteristics

steep sections to less steeper ones. This is due to the complete dissipation of the second electron cloud in a beam tetrode under the influence of a small anode voltage, while in a pentode the suppressor weakens the action of the anode (screens the electron cloud from the anode), and thus obstructs the dissipation of the cloud. The

dynatron effect is not fully eliminated in beam tetrodes and takes place at low anode currents, i.e., at considerable negative values of control grid voltage, when the space charge is of insufficient intensity to effectively "brake down" the secondary electrons. In this case a dip is formed in the characteristics. This part of the characteristics is not used, however, in practical applications.

The parameters of beam tetrodes are determined from their characteristics in the same way as in the case of pentodes. A triode connection of beam tetrodes is sometimes employed, the screen grid being connected to the anode. Approximate parameters of beam tetrodes are as follows: mutual conductance is within 1-20 ma/v, as in other types of valve, anode resistance is from several tens of thousands to hundreds of thousands of ohms, and amplification factor — from tens to hundreds.

45. VARIABLE-MU VALVES

The high amplification factor of pentodes employed in high-frequency amplifiers is useful only on reception of weak signals. Strong signals, such as those received from nearby radio stations, are badly distorted on such high amplification.

Requirements for convenient automatic adjustment of amplification, depending upon the strength of the signals, has led to the evolution of high-frequency amplification pentodes with a special form of characteristic, shown in Fig. 91*a*. The lower bend of such a characteristic is quite extended and is sometimes referred to as the "tail". In valves with this type of characteristic the control grid has variable density; the middle part is coarse, while the extreme parts of the grid are made finer (Fig. 91*b*). The effect of this is that of having, apparently, two valves in a common valve envelope, one supplied with a fine grid and the other with a coarse one. As we already know, the finer the grid, the lower value of grid voltage is required to cut off the anode current. In a valve of the described construction strong negative potential applied to the grid causes the fine extreme parts of the grid to create a condition of cutoff, only the coarse middle section of the grid passing and controlling a part of the anode current.

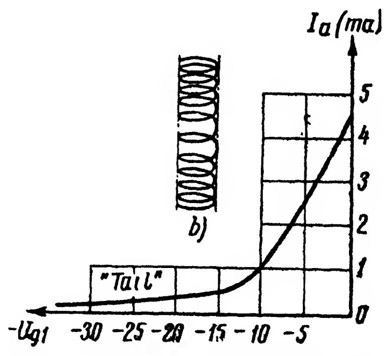


Fig. 91. Extended characteristics of a valve and the construction of the control grid of such a valve

This results in a characteristic noted for low value of mutual conductance S and low amplification factor μ, i.e., the valve operates on the bottom (the lower) part of the characteristic. On low negative grid potentials the whole area of the control grid operates, and in this case the coarse middle part of the grid plays only a secondary role, the main control being taken over by the fine extreme parts. This gives the valve high mutual conductance and high amplification factor.*

Valves with variable grid density are called variable-mu valves, in which the reception of weak signals is controlled on the steep part of characteristic, while the reception of strong signals is controlled in the section with smaller slope (the "tail" part of the curve). Since the "tail" part of the characteristic offers low amplification, such a valve gives distortionless reception of very strong signals. In circuit diagrams, variable-mu valves have the same schematic representation as ordinary valves. The application of variable-mu valves is discussed in greater detail in Chapter IX (Section 114).

46. RECEIVING AND LOW-POWER AMPLIFYING TETRODES AND PENTODES**

Beam tetrodes are designed for application in output stages of low-frequency amplifiers. The most popular filamentary beam tetrode types are 2П1П and 2П2П.

The most commonly used cathode-type beam tetrodes are represented by: 6П1П, 6П3С, 6П6С, 6П7С, 6П13С, and 30П1С. The last of these (30П1С) operates with normal anode voltage of 110 volts and is designed for receivers and amplifiers working from rectifiers without step-up transformers. In valve marking, letter П indicates low-frequency output pentodes and beam tetrodes.

In general any pentode may be classified as belonging to one of two groups: pentodes designed for high-frequency amplification, and more powerful pentodes intended for low-frequency amplification service.

The marking of *high-frequency pentodes* with normal characteristics includes letter Ж, standing after the first figure of the designation (for instance — 6Ж7). In variable-mu pentodes, letter К is used instead of Ж (for instance — 6К7).

In the past pentodes were so made that their control grid terminal was provided at the top of the valve envelope, while one of the valve base pins was used as anode terminal. Such a construction

* In modern valves employing the described principle the required variable density of various grid sections is calculated on the basis of complex formulas, and the transition of density is not as simple as Fig. 91*b* would indicate. In this way the most desirable shape of valve characteristic is obtained.

** Transmitting tetrodes and pentodes are studied in Chapter VIII.

had its advantages and disadvantages. The main advantage was the reduction of parasitic anode-grid capacitance. One of the disadvantages was wiring difficulties, because with this type of valve design it becomes necessary to use special shielded wires in order to connect the grid terminal (located on the top of the valve) to external circuit.

In this respect modern "single-ended" valves are more convenient, because in the new valves all the electrodes are brought out through the base. In order to obtain the smallest value of parasitic anode-

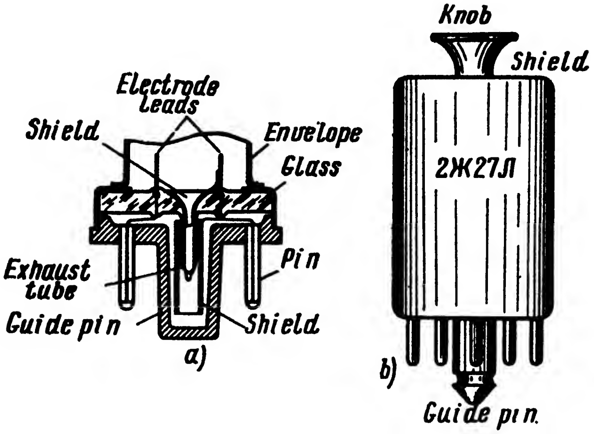


Fig. 92. Electrode terminations in a single-ended high-frequency valve *(a)* and external view of octal valve *(b)*

grid capacitance, anode and grid are connected to diametrically opposite pins of the valve base, while inside the base and in its guiding pin a metal shield is provided (Fig. 92*a*). As a result the parasitic anode-grid capacitance of the new valves is very small, too.

Below are listed various types of high-frequency pentodes widely used at present.

Bantam pentodes: 2Ж27П, 1К1П, 1К2П (filamentary valves); 6Ж1П, 6Ж2П, 6Ж3П, 6Ж4П, 6Ж5П, 6К1П, 6К4П (cathode-type valves).

Valves with control grid terminal brought out at the top: 6Ж7, 6Ж6С, 6К7, 6К9С (cathode-type valves).

Single-ended metal valves: 6Ж3, 6Ж4, 6Ж8, 6К3, 6К4, 12Ж8, 12К3, 12К4 (all these valves are of the cathode type).

Loctal valves: 2Ж27Л (filamentary valves); 4Ж1Л and 12Ж1Л (cathode-type valves).

Cathode-type miniature valves: 6Ж1Б and 6Ж2Б.

Cathode-type acorn valves: 6Ж1Ж and 6К1Ж.

Filamentary miniature valves: 2Ж2M and 2K2M.

Loctal pentodes (Fig. 92*b*) are provided with glass envelopes, usually contained in a solid metal shield. Eight pins, sealed in a thick glass bottom, serve as electrode terminals and pass through holes in the bottom of the shield. In this construction the guide pin, provided with a groove (lock), is an integral part of the shield. When a valve of this type is inserted in its socket, special springs snap into the groove on the guide pin. This firmly holds the valve in the socket even in the presence of mechanical shocks. The metal guide pin simultaneously serves as a shield, reducing the capacitance between anode and control grid terminals. A special knob located on the top of the valve facilitates the insertion and removal of the valve from its socket.

Miniature pentodes 2Ж2M and 2K2M are old-type valves. Their glass envelopes are coated with a layer of metal, serving as a shield and connected to pin *1* of the valve base. This shield eliminates parasitic capacitances between valve electrodes and external parts of the radio circuit.

All high-frequency pentodes are successfully used in low-frequency amplification circuits.

Low-frequency pentodes are manufactured for application in final stages of amplifiers, and all their terminals are brought out through the base without any shielding procautions, which are not needed on low frequencies where a small value of parasitic capacity C_{ag} can cause no undesirable effects. The following valves belong to this class: 1П2Б, 1П3Б, 1П4Б, 6П9, 6П15П, 6П18П, and 6Ф6С (the last one now outmoded). Previously manufactured miniature filamentary-type pentodes СБ-244 (or СО-244) and СБ-258 (or СО-258) are sometimes encountered but no longer made. Pentodes 06Ж6Б and 06П2Б are intended for low-frequency amplification.

47. COMPLEX VALVES

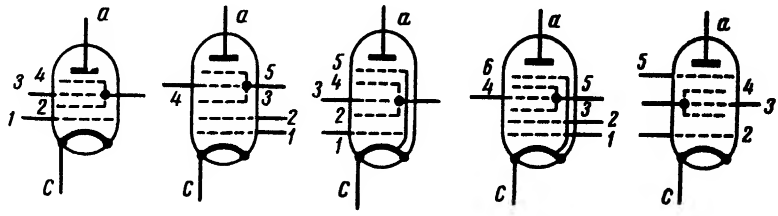
Apart from pentodes, radio equipment widely employs multi-grid and combined valves. Multi-grid valves are provided with two control grids to which alternating voltages of different frequencies are fed, giving them dual control over anode current. Applications of these valves are discussed in Chapter IX. Below are given, in brief, only their general features.

There are four types of complex multi-grid valves:

1. *Hexode.* This is a six-electrode valve with four grids (Fig. 93*a*). Grids *1* and *3* are control grids, while grids *2* and *4* are screen grids. Grid *4* operates as the usual screen grid of a tetrode, i.e., it serves to increase amplification factor μ and to decrease parasitic capacitance C_{ag1}. Grid *2* eliminates parasitic capacitance between the control grids.

2. *Heptode.* Heptode, otherwise called pentagrid, is a seven-electrode valve, five elements of which are grids. Valves of this type are used for frequency conversion in radio receivers, which is discussed in detail in Chapter IX. Letter *A* is used to denote frequency converter valves with two control grids.

Fig. 93*b* gives a schematic representation of the old-fashioned *heptode-converter*. Grids *1* and *2*, together with the cathode, work as a triode in the circuit of a low-power valve oscillator. Hence grid *1* is the control grid, while grid *2* serves as the anode of the oscillator. Grid *4* is also a control grid. Grids *3* and *5* are screen grids; grid *5*



a) b) c) d) e)

Fig. 93. Schematic representation of multi-grid valves: *a)* hexode; *b)* heptode-converter of old type; *c)* heptode-mixer; *d)* octode; *e)* heptode-converter of new type

acts as the usual screen grid, as in a tetrode, while grid *3* serves to eliminate the capacitive coupling between the two control grids. Such a heptode-converter acts like a tetrode with two auxiliary grids forming, together with the cathode, a triode. Besides this additional triode section, the valve also contains another screen grid whose function is to separate the triode and tetrode sections. Both sections utilise the common electron stream. Cathode-type valve 6A8 and valve СБ-242 (or СО-242) belong to this type. Miniature versions of valves designed for battery-operated sets are СБ-242 (or СО-242).

Production of improved heptode-converter valves has begun, including such valves as: 6A7, 6A10C and 6A2П (cathode-type valves) and filamentary 1A1П and 1A2П (Fig. 93*e*). In these valves grid *1* is the control grid of the triode section, while grid *2* simultaneously serves as a screen grid and the triode's anode. Grid *3* is the second control grid. Grid *4*, connected with grid *2* in the valve, is also a screen grid. Grid *5* is the suppressor grid. Another *heptode-mixer* (five-grid mixer) type 6Л7 used to be produced. In this type grids *1* and 3 are control grids, grids *2* and *4* — screen grids, and grid *5* is the suppressor grid, connected to cathode inside the valve (Fig. 93*c*).

3. *Octode.* This is an eight-electrode valve with six grids, shown schematically in Fig. 93*d*. It differs from the heptode-converter

described above (Fig. 93b) in that it has a sixth (suppressor) grid connected with the cathode.

4. *Nonode*. Nonode is the name of a seven-grid (nine-electrode) valve recently developed in Western Europe. It is designed for application as a limiter and detector of frequency-modulated signals in radio receivers.

The desirability of smaller valve dimensions, simplification of wiring and economy of power supply have led to the development of various combined valves which contain, in a common envelope, two and sometimes three or four valves, each one of which has its own electrodes.

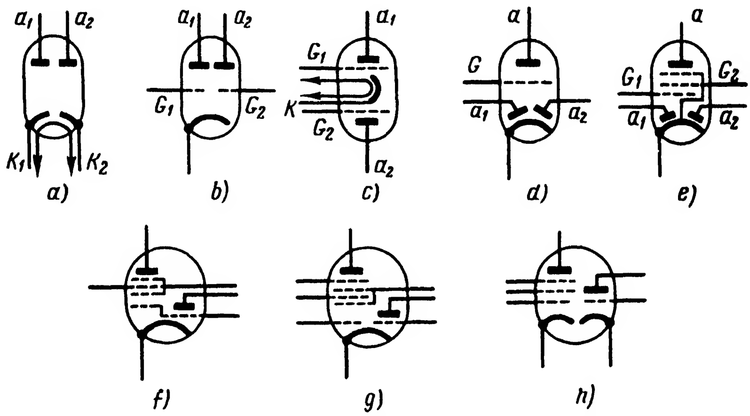


Fig. 94. Schematic representation of complex combined valves: *a)* double diode; *b) and c)* double triode; *d)* double diode-triode; *e)* double diode-pentode; *f)* triode-hexode; *g)* triode-heptode; *h)* triode-pentode

All the heater leads of these individual valves are connected within the envelope, and only two common heater terminals, together with the terminals of other valve electrodes, are brought out. The cathodes of such valves are either brought out separately or else connected to common pins. The heaters are usually connected in parallel. In some of these combined valves a common cathode is employed for various valve sections. For the sake of simplicity, schematic representations of combined valves frequently show only one heater and one cathode.

The following letters are used in marking various types of complex valves: X — double diode designed for detection purposes; H — double triode; Г — triode with one or several diodes; Б — pentode with one or several diodes; Ф — triode-pentode; И — triode-hexode or triode-heptode. Below is given a brief summary of complex valve types used in the Soviet Union.

Double diode. Double diodes are used as detectors in radio receivers. Fig. 94*a* gives a schematic representation of type 6X6C double diode. This type should not be confused with a twin-anode kenotron. The latter, although it also contains two diode elements, is designed exclusively for rectification of a.c. mains power. Valves 6X2П and 12X3C are two other types of double diodes.

Double triode. Figs 94*b* and 94*c* show two methods of schematic representation of double triodes. The most popular types are: 6H7C and 6H15П — with common cathode terminal; 6H1П, 6H2П, 6H3П, 6H4П, 6H5П, 6H5C, 6H8C, and 6H9C — with separate cathode terminals; filamentary double triode 1H3C, designed for economical operation of battery-type radio equipment. Old-type CO-243 double triode, designed for 2-volt operation, is sometimes encountered in practice. Double triodes are chiefly used in low-frequency amplifiers, but are also employed in some other circuits.

Triode with diodes. This is a combination of detector diodes with a triode, the latter being usually employed for low-frequency amplification. In most cases such valves contain two diodes (6Г7), the control grid terminal of the associated triode located at the top of the valve. Equally popular are single-ended valves 6Г1, 6Г2, 12Г1, 12Г2 (see Fig. 94*d* for schematic representation), and the triple diode-triode 6Г3П.

Pentode with one or two diodes. This type of valve is shown schematically in Fig. 94*e*. Here the diodes are usually employed for detection, while the pentode serves as a low-frequency amplifier, although it is sometimes used for intermediate-frequency amplification and other purposes. Double diode-pentode 6Б8C and diode-pentodes 1Б1П and 1Б2П belong to this type of valve.

Triode-hexode (Fig. 94*f*). Triode-hexode valves are used for frequency conversion (like heptode-converter) and are a combination of triode-oscillator and hexode.

Triode-heptode (Fig. 94*g*). Valves of this type are also employed for frequency conversion. At present the Soviet valve-manufacturing industry produces indirectly-heated triode-heptodes of bantam type 6И1П.

Triode-pentode (Fig. 94*h*). This is a combination of triode and pentode. One such valve now produced is known as 6Ф1П and has separate cathode terminals.

It should be noted that pentodes are sometimes used as frequency converters, and in such cases the suppressor acts as the second control grid. Pentodes are sometimes made to operate in the following way: the cathode, control grid and screen grid function as a triode, and the space between the cathode and anode as the interelectrode space of a diode.

48. INTERCHANGEABILITY OF VALVES

It often happens in practice that when a valve of required type is not available, a valve of another type has to be used in its place. This is a simple matter if the parameters of the substitute valve differ only a little from those of the original valve and if both types of valves have similar sockets and power supply rating. The problem becomes more difficult when the valves differ, for instance, in their socket connections. In such cases various contrivances and modifications have to be resorted to (transitory sockets, changes in socket connections, etc).

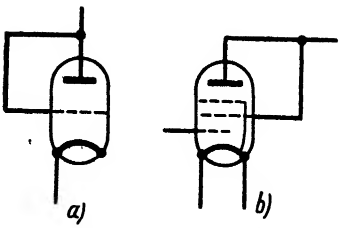


Fig. 95. Triode connected as a diode, and pentode connected as a triode

In solving such problems it should be remembered that in many cases a complex valve will successfully perform the functions of a simpler valve, providing power supply ratings of both valves are the same. Substitution, of course, calls for appropriate connections between valve electrodes. For instance, a triode can function as a diode, if its anode and grid are connected together (Fig. 95*a*). A tetrode or pentode can be used as a triode if the screen grid is connected to the anode (Fig. 95*b*).

Complex valves are sometimes used as simple ones in circuits where it is desirable to use one type of valve in all the stages. In such situations, if most valves are, for instance, of pentode type, and only one or two triodes are required in the circuit, it is wise to substitute the same type of pentodes for the required triodes, connecting the substitute valves as triodes.

A combined valve may be sometimes replaced with two simpler valves. For instance a diode-pentode can be replaced by a double-diode and a pentode.

49. VALVE TESTING

Of all parts of radio equipment, radio valves are the first to go out of order. Therefore the search for failure in radio equipment should always begin with valve testing. Special types of radio testers are available (such as ИЛ-12, ИЛ-14 and others), but the first thing to check when testing a radio valve for proper operation is that it is inserted in the right socket of correctly operated radio equipment. The operation of the equipment will then attest to the quality of the valve.

It may happen, however, that no such correctly functioning equipment is on hand. In such cases some means has to be devised for testing the valve without the help of equipment.

The ordinary tester (avometer) is used for checking the continuity of filaments and heaters. The same device is employed for locating short-circuits between electrodes. In lieu of an avometer, a primitive tester can be easily made up of a voltmeter and a battery (it is also possible to use a pair of earphones or a torchlight bulb instead of the voltmeter).

When a voltmeter-type tester is used, the procedure of valve testing is as follows. Connect the voltmeter into the filament circuit of the valve, as shown in Fig. 96*a*. If the meter gives a reading, the filament is intact. After this test check the valve for short-circuits between various electrodes. To do this connect the voltmeter between valve electrode terminals, which should not be connected to each other. The absence of voltmeter readings, in this case, shows that there are no inter-electrode short-circuits in the valve.

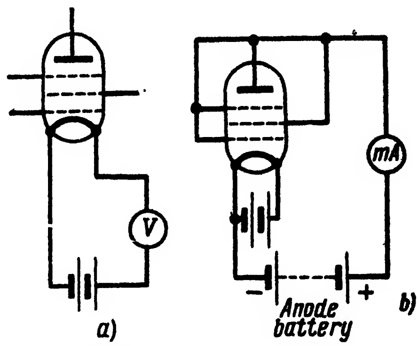


Fig. 96. Valve testing circuits for continuity of filament and presence of emission

The valve is tested for emission in the following way (Fig. 96*b*). Connect it as shown; in this circuit arrangement all the grids are connected to the anode, which makes the valve function as a diode. Apply normal voltage to the valve filament (or to the heater of a cathode-type valve). The anode voltage, however, should be considerably reduced and must not exceed 10-20 volts. Connect a milliammeter with a corresponding scale into the anode circuit of the valve. If the valve has not lost its emission, the meter will give a reading. To ascertain whether the emission is normal, connect a good valve in place of the one being tested and compare the emission current of both valves.

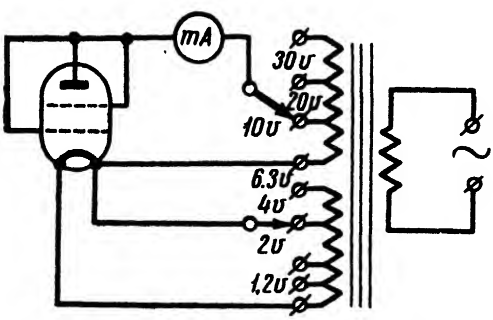


Fig. 97. Valve emission testing circuit operating from a.c. main

Note the reading obtained with the good valve and in future use it as a reference when testing any other valves of the same type.

A valve can be also tested for emission without the anode battery; it is merely sufficient to connect the anode circuit of the valve to the positive terminal of the filament battery. The anode current will be

very small in this case, however, and a low-scale milliammeter must be employed in such a test.

Another way of testing a valve is shown in Fig. 97. In this case no batteries of any kind are used. A transformer, connected to alternating-current mains, is used in place of filament and anode batteries. It must have a filament winding provided with taps to give several values of voltage (for instance: 1.2; 2; 4 and 6.3 v) for testing various types of valves. A separate anode winding, wound on the core of the same transformer, must be also provided with taps (10, 20 and 30 v). When assembling this simple valve tester it is worth while to install various types of valve sockets in it and to make provisions for the connection of anodes with grids.

50. CATHODE-RAY TUBES (CRT)

Cathode-ray tubes are widely used in various types of radio measurements, radiolocation, television and many other fields of radio and electronics. There are various types of cathode-ray tube. Let us first examine cathode-ray tubes employing electrostatic focusing and electrostatic deflection; such tubes are frequently called simply electrostatic tubes.

Figs 98*a* and 98*b* show such a tube, its supply circuits and circuit designation. Tube electrodes, each one of which has its own function, are placed in a cone-shaped envelope.

Indirectly-heated oxide-coated *cathode C* is a small metal cylinder containing *heater element HH*. Electrons are emitted by the oxide-coated bottom of the cathode. *Control electrode CE*, a cylinder with a hole in its bottom, is placed near the cathode. This electrode, also referred to as grid or modulator, is supplied with small negative voltage (several fractions of a volt) in respect to the cathode. The voltage is adjusted by means of potentiometer R_1. The electric field formed between the control electrode and the cathode contracts the electron stream emitted by the cathode and directs it into the hole of the control element. If the negative voltage of the control electrode is made to increase, a continually greater number of electrons will be repelled by this electrode back to the cathode, and a fewer number of electrons will be allowed to pass through the opening of the control electrode. At a certain negative voltage of the control electrode all the electrons will be repelled back to the cathode.

The other two electrodes of a cathode-ray tube are called the *first* and *second anodes* (a_1 and a_2). Both are of cylindrical design and are charged with high positive potential in respect to the cathode. Voltage U_2 of the second anode is from about 600 v to several thousands of volts, while voltage U_1 of the second anode is several times smaller. The first anode has partitions provided with holes (dia-

phragms). Certain cathode-ray tubes contain three or even more anodes of complex design.

The accelerating action of anode fields makes the electrons travel at a very high velocity and focuses them, with the help of the diaphragms, into a fine electron pencil, the *electron ray*. The whole system, consisting of cathode, control element and anodes, is designed for forming just such a ray and is called an *electron gun*.

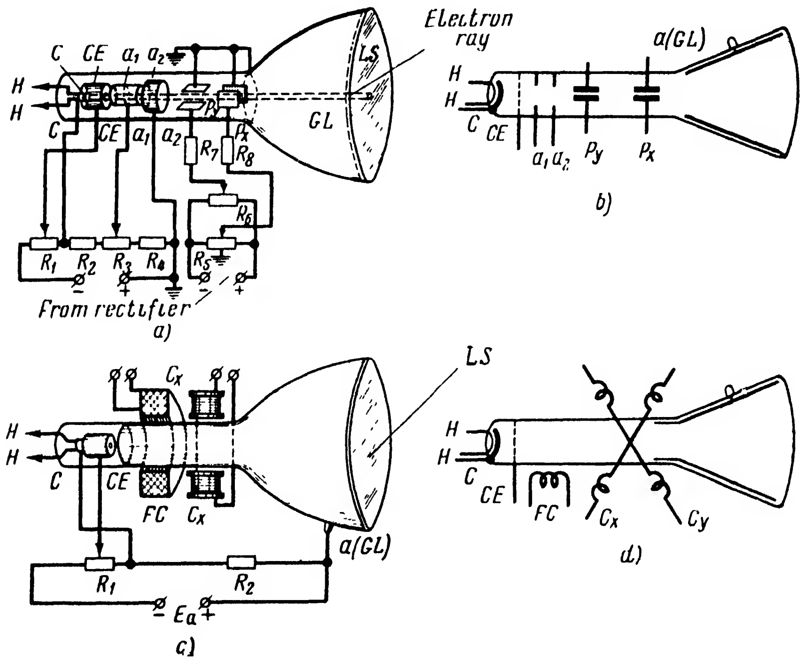


Fig. 98. Electrostatic *(a)* and electromagnetic *(c)* cathode-ray tubes and their schematic representation in circuit diagram *(b* and *d)*

The electron ray passes the whole length of the tube and strikes *luminescent screen LS*. This screen is the bottom glass part of the tube and is covered with a layer of chemical substance (e.g., zinc oxide, etc.) which glows under the impact of high-velocity electrons. A spot of light appears at the point where the electron ray hits the screen. Various types of chemical substances covering the inner part of the screen give different kinds of glow. Screens designed for visual observation glow green, while those designed for oscillogram photography glow blue.*

The screens of television tubes glow white.

* The eye is most sensitive to green glow, while the photographic film — to blue.

The brilliance of glow of the cathode-ray tube screen depends upon the quantity of electrons hitting such a screen. The brilliance is adjusted by varying the negative potential of the control element. In the circuit being studied, potentiometer R_1 is the ***brilliance control***. Varying the potential difference between the anodes focuses the electron-ray. Potentiometer R_3 is the *focus control*. This control varies the voltage of the first anode, and the changing field intensity between the anodes sharpens or broadens the electron-ray focusing as desired.

Two pairs of *deflecting plates* P_Y and P_X are installed at right angles to each other along the electron ray path. When no potential difference exists between these plates they do not influence the electron ray.

However, when voltage is applied between the two plates of any pair, and an electric field is created between the two plates, this field will deflect the ray towards positively-charged plate. The higher the voltage applied to the plates, the larger will be the ray deflection from its initial position and the greater will be the shifting of the spot on the screen.

Plates P_Y deflect the ray in the vertical direction and are therefore called vertical-deflection plates or simply *vertical plates*. The other two plates P_X deflect the ray horizontally and are called *horizontal plates*.

The second anode of the *CRT* is usually connected to the chassis and to the earth, i.e., is at zero potential in respect to the earth. At the same time the cathode of the tube is isolated from the chassis and is at a high negative potential.

It is therefore very dangerous to touch the cathode and heater wires.

Do not touch any part of the *CRT* circuit, unless the circuit is disconnected from its power supply.

Of each pair of deflecting plates one is usually connected to anode a_2 and, therefore, is also placed at zero potential in respect to earth.

The electron ray is very sensitive to external electric and magnetic fields. Hence *CRT*s are usually placed in shields, and the terminals of their electrodes are brought out to the pins of the tube base. For the sake of simplicity, Fig. 98 shows electrode terminals located in various parts of the tube.

Electrons which reach the screen must be led away from it, otherwise the screen will become charged to a high negative potential. To avoid this the screen is so made that it has a considerable secondary emission. Besides, a part of the inner surface of the tube is covered with conducting (graphite) layer *GL*, connected to earth. The secondary electrones dislodged from the screen (which is always at a certain negative potential when the tube operates) fly to this graphite layer because it is charged positively in respect to the screen.

The graphite layer is sometimes used as the third anode. In such application it can be supplied with higher voltage than the second anode.

For the purpose of setting the light spot to its initial position on the screen, a certain value of direct voltage is fed to potentiometers R_5 and R_6. The sliders of these potentiometers are connected through high-value resistors R_7 and R_8 to the unearthed deflecting plates. When these sliders are set to middle positions, the deflecting plates' voltage is zero. Moving the sliders to the right or left applies either positive or negative voltage to the deflecting plates, which shifts the light spot in vertical or horizontal directions, as needed.

The electron gun is supplied with power from a rectifier terminated with voltage divider R_1, R_2, R_3, R_4. Various required voltages are taken off this voltage divider to appropriate electrodes of the electron gun. The current consumed by the tube usually does not exceed a fraction of one milliampere, and the divider also draws small current. Hence the rectifier is designed to give a high-voltage low-current output. The heater of the electron gun generally operates on alternating current.

The basic specifications of the *CPT* comprise power supply ratings, heater current, screen diameter and sensitivity.

The sensitivity of a cathode-ray tube is the deflection of the light spot in millimetres when the deflecting plate voltage changes by one volt.

This sensitivity is expressed in millimetres per volt (mm/v). The sensitivity of modern cathode-ray tubes is from 0.1 to 0.5 mm/v, and it is not the same for the vertical and horizontal plates. The plates farthest from the screen give the highest sensitivity. The sensitivity of the *CRT* decreases with an increase of voltage applied to the second anode.

The electron ray has a negligible inertia and obediently follows the voltage changes of the deflecting plates, even if such changes occur millions of times per second.

Some cathode-ray tubes employ electromagnetic focusing and deflection of the electron ray. A tube of this type, called an electromagnetic cathode-ray tube, is shown in Fig. 98 *c* and *d*. The electron gun of such a tube has only one anode. The graphite layer is used as the second anode, and in some tubes it is the only anode.

The control electrode and the anode are supplied with power in the same way as in an electrostatic cathode-ray tube. The slightly diverging electron stream emitted by the gun enters the longitudinal magnetic field created by focusing coil *FC* (Fig. 98). This coil is supplied with direct current. Under the influence of the magnetic field of the focusing coil the magnetic flux is focused, i.e., moves to the luminescent screen in a converging pencil. The action of focusing coil *FC* can be increased or decreased by changing the current flowing through the coil.

Two pairs of deflecting coils located at right angles to each other are used for the deflection of the ray. One pair, located vertically, is shown in Fig. 98*c*. Coils C_X, creating a field with the vertical direction of magnetic lines of force, deflect the ray horizontally. Coils C_Y deflect it vertically, being located horizontally and creating a field with horizontal magnetic lines of force.

The greater the number of turns of deflecting coils and the greater the current passing through them, the greater is the deflection of the electron ray. The direction of its deflection depends upon the direction of current flowing through deflection coils.

The sensitivity of electromagnetic tubes is given as the ratio of light-spot deflection to ampere-turns of respective pair of deflecting coils. The sensitivity of such a tube is, accordingly, measured in millimetres per ampere-turn.

In comparison with the electrostatic tubes, the electromagnetic tubes have a simpler design of electrodes, offer a better focusing of the electron ray and are of smaller length. However, they have a disadvantage of constant current consumption by their deflecting coils. The deflecting plates of an electrostatic tube draw no current; such tubes require only application of voltage to their plates.

Cathode-ray tube marking begins with a figure denoting the diameter or the diagonal of the screen in centimetres. Then come two letters, ЛО, denoting oscillographic and receiving television tubes (picture tubes) with electrostatic deflection of the beam; ЛМ, denoting oscillographic tubes with magnetic deflection, and ЛК, denoting kinescopes with magnetic deflection. The next marking symbol is a figure serving to differentiate between various tubes whose other marking elements are the same. In certain tubes the marking formula ends with a letter showing the type of luminescent screen (e.g., letter Б stands for white glow).

Among cathode-ray tubes the most popular types of oscillographic tubes are: 5ЛО38, 7ЛО55, 8ЛО29, 10ЛО43, 13ЛО36, 13ЛО37, 13ЛО48, 13ЛО54, 18ЛО33 and 31ЛО33. The most popular kinescopes are: 18ЛК4Б, 18ЛК5Б, 18ЛК15, 18ЛО40Б, 23ЛК1Б, 31ЛК2Б, 35ЛК2Б, 40ЛК1Б, 43ЛК2Б and 53ЛК2Б. The last three types of tube have a metal-glass envelope, while kinescopes 35ЛК2Б, 43ЛК2Б and 53ЛК2Б have a rectangular screen, electrostatic focusing and electromagnetic deflection.

51. NEON LAMP

Apart from electron valves, whose operation is based on passing electron streams through the vacuum, so-called *ionic devices* have also found wide application. Like the electrodes of an electron valve, the electrodes of ionic devices are also enclosed in envelopes, but the latter contain a low-pressure gas, owing to the presence of which

the process of ionisation takes place. This considerably complicates operation of the valve and changes its properties.

Ionic devices with the so-called *glow-discharge* action include *luminous gas lamps.* Representative of such lamps are *the neon* lamps which are widely used in radio engineering.

Neon lamps are employed as high-frequency indicators in tuned circuits of electron valve oscillators, in transmitter aerial circuits, etc., i.e., they can function as indicators when adjusting a tuned circuit to resonance, serve as locators of high voltage, etc.

The external view and schematic representation of a neon lamp are given in Fig. 99. The envelope of a neon lamp contains low-pressure neon gas (several millimetres or several dozens of millimetres of the mercury column) and two steel or aluminium electrodes shaped as plates, cylinders or wires. Two steel or aluminium electrode terminals are brought out to the lamp base or to special contacts.

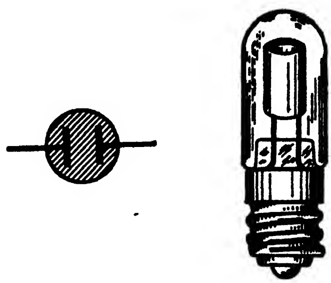


Fig. 99. Schematic representation and external view of a neon lamp

A neon lamp has no filament nor heated cathode. If the voltage across a neon lamp is less than a certain value (predetermined for different types of neon lamps and called *firing voltage)*, the lamp conducts no current. However, when the firing voltage is reached, ionisation of the gas will take place and the lamp will begin to conduct. The conduction of current by a neon lamp is accompanied by an orange-reddish glow, the intensity of which increases with the increase of voltage.

On direct current the glow chiefly takes place around the cathode; on alternating voltage both electrodes glow. If the voltage across a neon lamp is decreased, at a certain voltage, called the *extinguishing voltage* (somewhat lower than the firing voltage), the neon lamp will go out and will stop passing current. A neon lamp conducts electric current by virtue of the positive ions dislodging electrons upon hitting the cathode. This type emission is typical for glow-discharge ionic devices.

The firing voltage of various neon lamps ranges from several tens to several hundreds of volts, and the current they pass ranges from fractions of one milliampere to several scores of milliamperes. A limiting resistor is always connected in series with a neon lamp, preventing the latter from damage. Such a resistor is frequently built into a neon lamp base.

The current passing through a neon lamp depends upon the value of the limiting resistor and the voltage of the power supply source.

52. QUESTIONS AND PROBLEMS

1. What explains the directional conductance of a diode?

2. In which case will the thermionic emission of a cathode in a diode valve be greater—when the anode voltage U_a is 10 volts or 50 volts?

3. Draw a circuit diagram showing three diodes with parallel-connected filaments, a common filament rheostat and a voltmeter, measuring the filament voltage.

4. What will happen to an electron emitted by a hot cathode if the anode of the valve is charged negatively in respect to the cathode?

5. Given a 6-volt filament battery, which is to supply with power a valve filament rated at 4 v and 80 ma; calculate the value of dropping resistor to be connected in series with the filament circuit.

6. What are advantages and disadvantages of activated cathodes?

7. Why does a small anode current flow in a valve when $U_a = 0$?

8. Certain valves continue to operate for several seconds after being switched off. What kind are they?

9. Why is it not permissible to operate many types of valves at the saturation value of the anode current when normal voltage is applied to their anodes?

10. What happens to the electron cloud in the vicinity of a cathode when the cathode temperature is changed?

11. The mutual conductance of a triode is given by $S = 2.5$ ma/v. What will be the anode current change, expressed in milliamperes, if the grid voltage is changed from -2 v to $+3$ v?

12. Changing the control grid voltage in a certain valve from -10 v to $+12$ v varies the anode current from its zero value up to the saturation value of 55 ma. Are these data sufficient for determining the mutual conductance of the valve?

13. The anode resistance of a valve is given by $R_i = 8{,}000$ ohms. What will be the anode current change in such a valve if the anode voltage is changed from 120 v to 100 v?

14. How does a grid voltage change affect the direct-current anode resistance of a valve?

15. Does the thermionic emission of a cathode depend upon the control grid voltage?

16. What role does anode resistance R_a play in an amplifier stage?

17. The amplification factor of a stage is given by $k = 25$. Alternating voltage applied to the control grid is equal to 0.1 v. The value of an anode load resistor is given by $R_a = 20{,}000$ ohms. Find the value of alternating voltage across the load resistor R_a and determine the value of alternating current flowing through R_a.

18. If the control grid of a valve is not connected to anything, it becomes charged up to a certain negative potential during operation of the valve. What explains this phenomenon?

19. The amplification factor of a valve is given by $\mu = 40$. A 2-volt change of control grid voltage results in a 6-milliampere anode current change, the anode voltage remaining constant. In such a valve, if the grid voltage is kept constant, what must be the anode voltage change to vary the anode current by 12 milliamperes?

20. On normal voltage applied to the anode of a triode, the characteristic of the given triode is situated in the region of negative control grid voltages ("left-handed" characteristic). What kind of grid — coarse or fine — has this triode? Is its amplification factor μ low or high?

21. The capacitance between anode and grid (C_{ag}) of a triode is 10 pf. Find the capacitive reactance existing between the anode and the grid of the given triode for the following frequencies: $f_1 = 4$ mc and $f_2 = 40$ cps.

22. Why is it that the voltage applied to the screen grid of a tetrode must be lower than the anode voltage?

23. Given a 240-volt anode power supply, if the screen grid of a valve is connected to such power supply through a 50,000-ohm dropping resistor and the screen-grid current is equal to 3 ma, find the voltage applied to the screen grid.

24. The filament of a valve conducts current: does this necessarily mean that the valve is in good operating condition?

25. Why is it that when a valve is being checked for emission and its anode is connected to all its grids, the test voltage applied to the anode must be below its normal value?

26. How is the secondary emission suppressed in beam tetrodes?

27. Draw two grid characteristics for various anode voltages and explain how the parameters μ, R_i and S can be found from such characteristics.

28. If the control grid stops 98% of the electric lines of force of the anode field, what is the amplification factor of the valve?

29. What is the dynamic operating condition of a valve and how does it differ from the static operating condition?

30. Find the amplification factor of a stage if the grid of the valve it employs is supplied with 150-millivolt alternating signal and this results in the appearance of a 3-volt alternating signal across the anode load resistor.

31. Why is it that a potentiometer, not a rheostat, must be used for anode voltage adjustment in a circuit employed for plotting valve characteristics?

32. A valve filament, operated at normal voltage of 4 v, consumes 80 ma. Calculate the resistance of the filament and explain why on direct measurement of this resistance with an ohmmeter the meter reads 5 ohms?

33. What will be the voltage of a control grid if the grid is connected to the negative terminal of the filament?

34. Draw a circuit diagram showing the connection of a pentode into a high-frequency amplifier stage, the screen-grid voltage being supplied from a voltage divider.

35. In an amplifier stage (with the anode load resistor connected) the anode voltage changes when the control grid voltage is varied. How will the anode voltage change if the control grid is made positive and if it is made negative?

36. A valve has the following parameters: $R_i = 20{,}000$ ohms; $\mu = 30$. Load resistance $R_a = 40{,}000$ ohms. Find the alternating voltage value across R_a and determine the amplification factor of the stage if alternating voltage applied to the control grid is equal to 1.2 v.

37. A 2-milliampere anode current change is obtained in a certain valve either when the anode voltage is changed by 50 v or the control grid voltage is varied by 1.25 v. Find parameters μ, R_i and S of the given valve.

CHAPTER V

RECTIFIERS

53. KENOTRON RECTIFIER CIRCUITS

Diode valves have found an important application as rectifiers, converting a.c. (alternating current) into d.c. (direct current) for the purpose of supplying various types of radio and electronic equipment.

Diodes functioning as rectifiers are called kenotrons, and the circuits in which they are thus employed are known as kenotron rectifier circuits, or simply kenotron rectifiers.

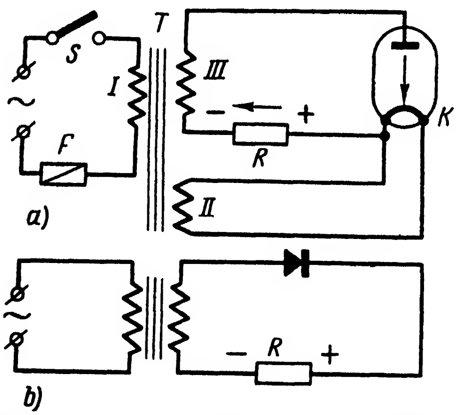


Fig. 100. The circuit of a half-wave rectifier

Kenotron rectifiers are used mainly for supplying d.c. power to the anode circuits of radio equipment, and in this respect they replace anode batteries. Besides kenotron rectifiers, other types of rectifiers can be also used for the same purpose.

The simplest circuit diagram of a kenotron rectifier is given in Fig. 100*a*. Primary winding *I* of power transformer *T* is connected to a.c. mains through switch *S* and fuse *F*. This transformer is provided with two secondary windings *II* and *III*, the first of which is a step-down winding, feeding the filament of the rectifier valve, while the other is a step-up winding.

Here pulsating current flows through load resistor *R*, while the filament of the kenotron valve acts as the positive terminal of this rectified pulsating current. The rectification process was already explained in Chapter IV (see Fig. 61). In this circuit only one half of each cycle is rectified, and such rectifiers are accordingly called *half-wave rectifiers*.

In rectifiers the filament circuit is considered as an auxiliary circuit, and sometimes it is not even shown. Circuit diagrams often show rectifier valves conventionally as rectifying units (Fig. 100*b*). The symbol is used for all types of rectifying units. The apex of the triangle bearing against the plate conventionally indicates the direction of current (electrons flow in the opposite direction).

Besides half-wave rectifiers, *full-wave rectifiers* are also used. These pass both halves of each cycle in one direction and produce a pulsating-current output (Fig. 101).

In the circuit diagram of Fig. 101*a* two separate kenotrons are employed, while circuit diagram of Fig. 101*b* pertains to a twin-anode kenotron. The step-up winding of the transformer has a centre-tap and must be designed to supply twice the voltage developed by the step-up winding of the half-wave rectifier. The two kenotrons or the two halves of the twin-anode kenotron function alternately. During the first half of a cycle the rectified current is passed by one half of the step-up winding of the kenotron whose anode is positive. The direction of this current is shown by arrow *1*. Then the polarity alternates, the anode of the other kenotron becomes positive, and this kenotron will pass the rectified current, which will flow through the other half of the step-up winding, as shown by arrow *2*.

In both cases currents passed alternately by each kenotron flow in the same direction and create the resultant pulsating

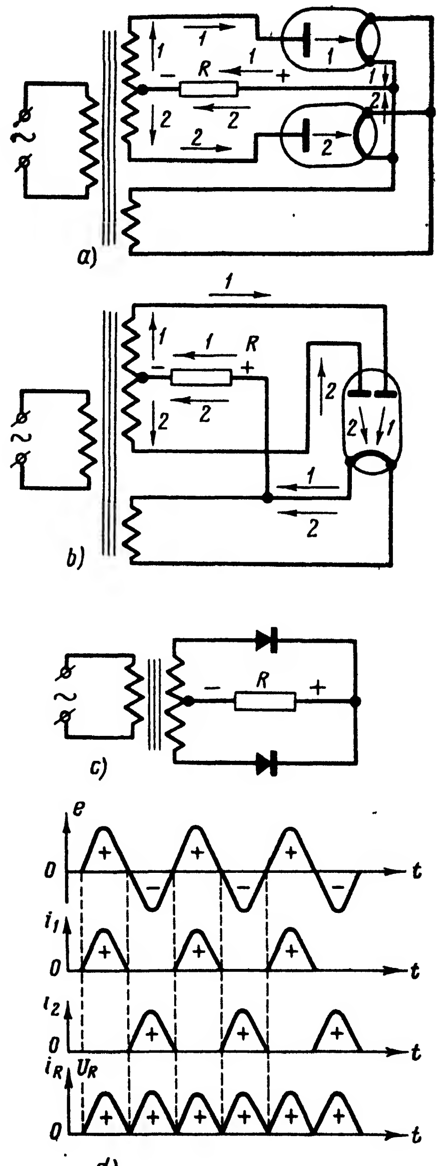


Fig. 101. The circuits of full-wave rectifiers and schematic representation of their rectification process

current, which flows through load resistor R and is represented by the lower curve in Fig. 101*d*. Here the first curve represents the voltage of the secondary winding of the transformer, while the second and the third show the currents rectified by each rectifying element. It would be correct to refer to the circuits shown in Fig. 101 as to *two-phase* circuits, because the step-up winding of the transformer has a centre-tap and hence acts as a source of two-phase e.m.f., i.e., it gives two electromotive forces equal in value but opposite in phase (with a phase shift of 180°).

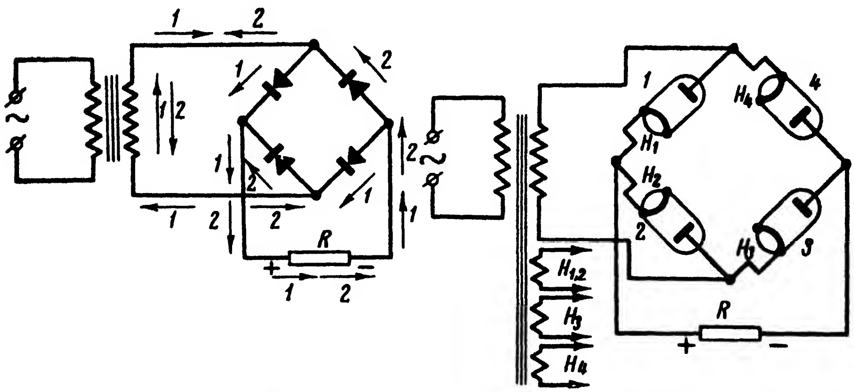


Fig. 102. Bridge rectifier circuit

Fig. 102 shows a bridge rectifier circuit. This gives full-wave rectification, although the step-up winding has no centre-tap and is designed for the same value of voltage as that of the simple half-wave rectifier. The disadvantage of the bridge circuit is that it has to employ four kenotrons, only two of which can use a common filament winding, the filaments of the other two kenotrons being supplied from two independent windings, thoroughly insulated from each other. Arrows *1* show the passage of current during the first half of a cycle, arrows *2* — during the second half of the cycle. The current flows through two rectifying units in series, and hence in such a circuit the voltage drop in the internal resistance of the rectifying units will be twice as great as in the rectifier shown in Fig. 101. This is another disadvantage of the bridge rectifier.

The circuit shown in Fig. 103*a* is of special interest. The transformer is not provided with a centre-tap double-voltage step-up winding, but the circuit offers full-wave rectification and simultaneous voltage doubling. This is a modification of the bridge circuit, in which two of the kenotrons have been replaced by capacitors. The two kenotrons left in the circuit must have separate filament windings, well insulated from each other.

In the first half of a period capacitor C_1 is charged (through kenotron K_1 whose anode is positive at the moment) up to the peak voltage value of the secondary winding. In the second half the anode of the other kenotron K_2 will be positive, and capacitor C_2 will be charged up through this kenotron almost up to the peak value of the secondary transformer winding. Capacitors C_1 and C_2 are connected in series, and consequently the resultant voltage across them is approximately equal to twice the peak voltage value of the transformer winding. Simultaneously with charging of capacitors

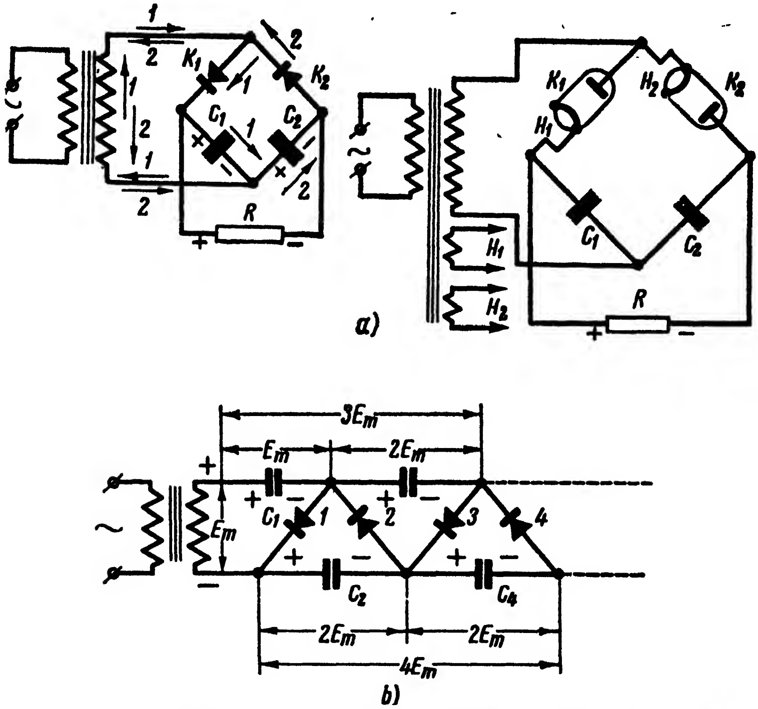


Fig. 103. Voltage-doubling *(a)* and voltage-quadrupling *(b)* rectifier circuits

C_1 and C_2 through kenotrons K_1 and K_2, these capacitors discharge through load resistor R. This lowers the voltage across the capacitors. The lower the resistance value of load resistor R (i.e., the greater the load current) and the smaller the capacitance of capacitors C_1 and C_2, the faster will the capacitors discharge and the lower will be the voltage across them. Hence it is actually impossible to obtain true voltage doubling with such a circuit; if the capacitance value of the capacitors is at least 10 mfd and the load current does not exceed 100 milliamperes, the output voltage will not exceed the transformer voltage by 1.7-1.9 times.

The advantage of the circuit just described is that the capacitors provide a certain amount of smoothing of the rectified current. This type of circuit, used together with a special kenotron provided with two anodes and two separate indirectly heated cathodes, is shown in Fig. 104. Such a kenotron needs a transformer with only one heater winding and is sometimes called a "doubling kenotron". Such a circuit, however, cannot be used when the rectified voltage value is to exceed 200-300 volts, because on higher voltages the insulation between the heater and cathodes can fail in the kenotron.

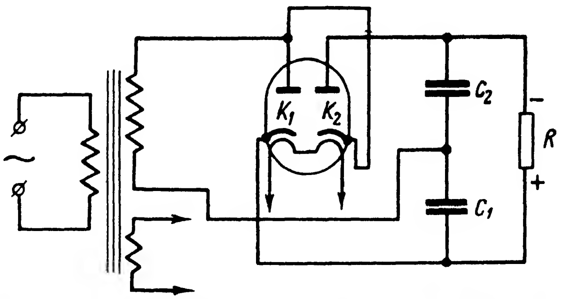


Fig. 104. Double-anode kenotron voltage-doubling circuit

It is possible to devise rectifier circuits which simultaneously multiply voltage any number of times. One of such circuits is given in Fig. 103*b*. This circuit quadruples the voltage and employs four rectifying units and four capacitors. During odd half-periods (for which the polarity of transformer voltage is shown), capacitor C_1 is charged through rectifying unit *1* almost up to the peak voltage value E_m of the transformer winding. Capacitor C_1, when charged, becomes a source of power supply. Therefore, during even half-periods, when the polarity of transformer voltage is changed, capacitor C_2 is charged up through rectifying unit *2* up to approximately doubled voltage value $2E_m$. This is the maximum value of resultant voltage developed by series-connected transformer and capacitor C_1.

In a similar manner during odd half-periods capacitor C_3 is charged through rectifying unit *3* also up to voltage $2E_m$, this voltage being the maximum resultant voltage developed by series-connected C_1, transformer and C_2 (it should be noted here that voltages appearing across C_1 and C_2 are directed opposite each other).

Continuing to analyse the circuit in this way, it can be concluded that capacitor C_4 will be charged up through rectifying element *4* again to voltage $2E_m$, which is the sum of the voltages across C_1 and C_3, transformer and C_2. Of course, the capacitors are charged to indicated voltage only gradually, i.e., several half-periods after the circuit has been switched on. As a result, a quadrupled voltage $4E_m$ will be developed across series-connected capacitors C_2 and C_4. Simultaneously, a tripled voltage $3E_m$ will be developed across capacitors C_1 and C_3. If additional capacitors and rectifying units are added to the circuit and connected in accordance with the same principle, then from the series of capacitors C_1, C_3, C_5, etc., it will be possible to obtain voltages multiplied an odd number of times (3, 5, 7, etc.), while capacitor series C_2, C_4, C_6, etc., will develop voltages multiplied an even number of times (2, 4, 6, etc).

When a load resistor is connected to one of the above voltages, the capacitors will begin to discharge and the voltage across them will be lowered. The lower

the resistance value of such a resistor, the faster will the capacitors discharge and the lower will become the voltage across them. Hence when the load resistance is not sufficiently high, such circuits become impractical. Actually circuits of this type can provide effective voltage multiplication only when the load current does not exceed 10-20 ma. Of course, it is possible to obtain greater values of current by increasing the capacitance of the capacitors, but an excessive increase is usually not worth while. The main advantage offered by such circuits is that of obtaining high voltage without a high-tension transformer. Another advantage is that the capacitors used by them need not be rated at higher voltages than $2E_m$, no matter how many times the voltage is multiplied, and each rectifying unit operates at maximum inverse voltage of only $2E_m$. If the rectifying units have a cathode requiring heating (e.g. — kenotron type rectifying units), there should be a separate winding for the heater of each one. It is therefore more convenient to use semiconductor rectifying units.

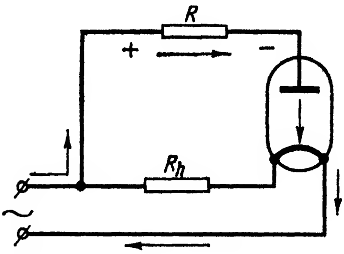


Fig. 105. Transformerless rectifier circuit

Small radio receivers sometimes use half-wave rectifiers requiring no power transformer and designed to operate directly from the mains (Fig. 105). In this case the heater circuit of the kenotron is connected to the mains through dropping resistor R_h, calculated from the following equation:

$$R_h = \frac{U - U_h}{I_h},$$

where

R_h — ohmic resistance value of dropping resistor;
U — mains voltage;
U_h — heater voltage rating of the kenotron;
I_h — kenotron heater current rating.

In this circuit the rectified voltage appears across load resistor R. The rectified current path is shown by the arrow.

54. SMOOTHING FILTERS

The current at the output of a rectifier has a pulsating character, i.e., it contains a.c. and d.c. components. A rectifier is intended to feed some type of load (e.g., anode circuits of receivers) with direct current; hence the a.c. component at the output of a rectifier is undesirable and should be kept away from the load. The d.c. component is the only part of the pulsating current that should be allowed to reach the load, and the value of this component should be as high as possible.

The pulsations of a rectified current are not sinusoidal; hence the a.c. component contains a number of harmonics. The first harmonic, whose amplitude is denoted by I_{m1}, is the strongest. In a

half-wave rectifier the direct current component (I_-) or direct voltage component (E_-) is only 0.32, i.e., approximately only one-third of the maximum value (Fig. 106), while the amplitude of the first harmonic of the alternating component is equal to 0.5 I_{max}.

The frequency of the a.c. component is equal to the frequency of the alternating current supply (50 cps on usual a.c. mains). As may be seen from above, the a.c. component is greater than the d.c. component, and this is a disadvantage of half-way rectifiers.

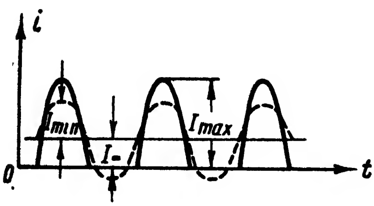


Fig. 106. Rectified current components of half-wave rectification

In any type of full-wave rectifier (with the exception of a voltage-doubling rectifier), $I_= \approx 0.64\ I_{max}$, and $I_{m1} \approx 0.42\ I_{max}$. The same relations also hold true for the voltages. The frequency of pulsations is doubled (100 cps on the usual a.c. mains). Apparently the d.c. component is greater than the a.c. component at the output of a full-wave rectifier. This is the advantage of full-wave over half-wave rectification.

In order to obtain really direct current and voltage in the load resistance, a filter must be employed to smooth out current pulsations. Such a filter must be connected between the rectifier and its load.

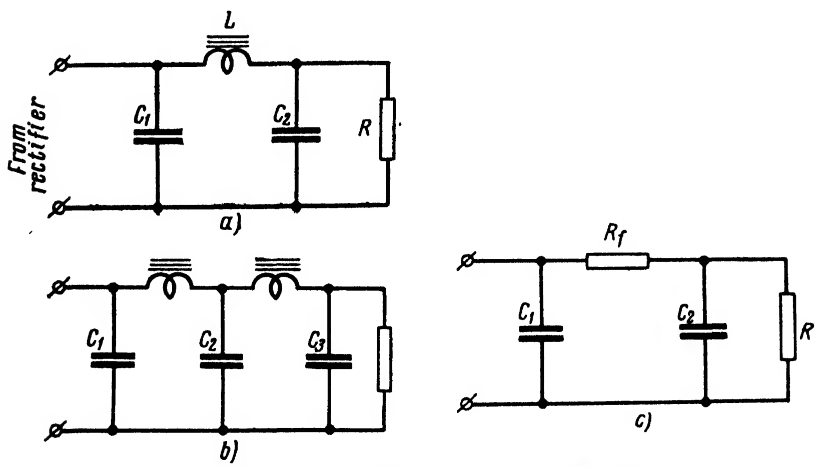


Fig. 107. Various circuits of smoothing filters

The function of a filter is that of passing d.c. component to the load and blocking the a.c. component.

Circuit diagrams of smoothing filters are given in Fig. 107. The capacitors used in smoothing filters have a capacitance value from

several microfarads to several dozens of microfarads. Electrolytic capacitors are widely used in such filters. The chokes are wound on iron cores, and their windings consist of several thousand turns to obtain inductance values ranging from several henries to dozens of henries.

A.c. components (of rectified currents), possessing frequencies of 50 cps, 100 cps and higher, are by-passed by capacitor C_1 (because of its low reactive capacitance), and return to the rectifier. At the same time the d.c. component easily passes through the choke and continues on its way to the load (the d.c. resistance of choke winding is low and there is only a small voltage drop across it, as far as the d.c. component is concerned, but the reactance of the high inductive choke L blocks the a.c. component).

It so happens, however, that a certain part of the a.c. component manages to pass through the choke. To suppress this undesirable component completely, an additional capacitor C_2 is connected in parallel with the load resistance. The capacitive reactance of this capacitor is small in comparison with the resistance of the load and, as a result, the greater part of the stray a.c. component, having passed through the choke, will be by-passed back to rectifier by capacitor C_2 and will not reach the load. The higher the inductive reactance of the choke and the smaller the capacitive reactance of the capacitor, the more effective will be the smoothing action of the filter. Thus the values of L and C should be as large as possible, because on this depends the degree of smoothing of the pulsations.

It is very significant that *the first capacitor (C_1) of the filter not only takes part in the smoothing prosess, but also considerably increases the d.c. component of the rectified voltage*. This effect is due to fast charging of C_1 by the rectifier when a voltage increase reaches it (during a pulse) and due to the following slow discharge of the same capacitor through the choke and through the load resistance R. The capacitor cannot discharge through the kenotron, because the electrons cannot pass from the anode to the cathode. The higher the value of load resistance R and of inductive reactance of the choke (opposing fast increase of current and fast decreases of voltage across the capacitor), the slower will the capacitor discharge. Therefore the rectified voltage is given no time to decrease, as it would do in the absence of C_1. Hence between pulses the voltage across C_1 begins to drop only slightly, and the next pulse delivered by the kenotron again brings up the charge of the capacitor to its previous value.

In Fig. 108 the solid line shows voltage change across capacitor C_1 in a half-wave rectifier. For the purpose of comparison, broken lines in the same drawing indicate pulse curves obtained in the absence of filter. As may be seen, the filter has considerably smoothed out the pulses and increased the level of d.c. component, which can be as high as 0.8-0.9 of a.c. peak voltage U_{max}.

The higher the value of the load resistance, the lower will be the discharge current of capacitor C_1 and the slower will be the decrease of voltage across it, i.e., the higher will be the value of d.c. component and the better the pulsation will be smoothed out.

If the load resistance is completely disconnected from the filter, capacitor C_1 will be charged up to the value of peak voltage U_{max}, equal to the maximum amplitude of transformer e.m.f. In such a case C_1 will not discharge at all, and the voltage across it will be perfectly stable and equal to U_{max}.

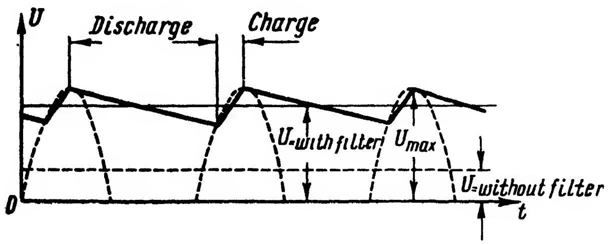


Fig. 108. Smoothing action of the first filter capacitor

Thus when a filter is connected to the output of a rectifier, the value of direct voltage delivered by the rectifier can even exceed the effective voltage value of the transformer and approach the transformer voltage peak value.

For instance, if the effective value of transformer voltage is 300 volts (as measured by a voltmeter), the peak value of the same voltage is equal to $1.4 \times 300 = 420$ v. If the d.c. component across capacitor C_1 is equal to $0.8\ U_{max}$, then $U_{=} = 0.8 \times 420 = 336$ v, which is higher than the effective value of voltage delivered by the transformer. In practice the d.c. voltage across the load resistance R will be lower than the above value because of voltage drop in the internal resistance of the kenotron and in choke winding.

Pulsating voltage, which is still present across capacitor C_1, is fed to choke L and capacitor C_2, which are connected in series and play the role of a voltage divider (Fig. 107*a*). The inductive reactance of the choke is much higher than the capacitive reactance of capacitor C_2. Therefore the choke will represent the higher resistance part of such a voltage divider. Consequently, the greater part of the pulsating voltage will appear across the choke, and only a small part of this voltage will appear across capacitor C_2 and load resistance R, which is connected in parallel with this capacitor.

The higher the pulse frequency, the better works the smoothing filter, because at higher frequencies the inductive reactance of the choke increases and the capacitive reactance of filter capacitors decreases. This is why it is difficult to smooth rectified voltage at the output of a half-wave rectifier, where the frequency of pulses is

50 cps, as compared with 100 cps at the output of a full-wave rectifier, where good smoothing is achieved easier.

We have studied above the function of a filter consisting of one section. When still better filtering is needed, filters consisting of several sections are sometimes used. An example of a twin-section filter is shown in Fig. 107*b*. Here capacitor C_1, the choke and capacitor C_2 represent one section of the filter. At the same time capacitor C_2 plays the role of input capacitor of the second section of the filter, which also includes another choke L_2 and capacitor C_3. This second filter section operates in the same way as the first section and provides additional filtering (smoothing) action.

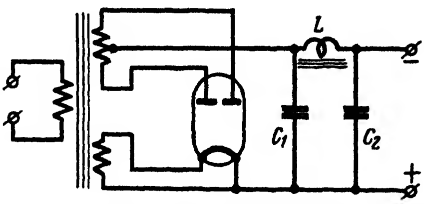


Fig. 109. The circuit diagram of a kenotron rectifier with a filter

When a rectified current is very small and a certain amount of voltage drop in the filter is tolerable, a resistance can be used instead of the choke for the purpose of making the filter cheaper and simpler. Such a filter is shown in Fig. 107*c*, where the smoothing resistor R_f can have a value from several thousand ohms to several tens of thousands of ohms.

Fig. 109 gives a circuit diagram of the most popular type of full-wave rectifier supplied with a filter.

The external characteristic of a rectifier is used when it is required to determine $U_=$ given by a rectifier at a certain load current $I_=$. Such characteristics are shown in Fig. 110 and pertain to a half-wave rectifier operating without a filter (curve *1*) and also operating with a filter provided with capacitor C_1 at its input (curve *2*). As may be seen from these characteristics, when no current is drawn by the load from rectifier ($I_= = 0$), $U_= = 0.32\ E_m$, if no capacitor is present at the output of the rectifier. When the capacitor is connected $U_=$ becomes equal to E_m, where E_m stands for the peak value of transformer voltage.

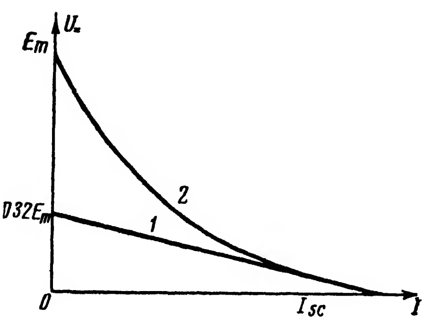


Fig. 110. External characteristics of half-wave rectifier operating without a filter *(1)*, and with a capacitor-input filter *(2)*

The lowering of the voltage upon current increase is attributed to the increase of voltage drop in the external resistance of the rectifier, this resistance being made up of the internal resistance of the

kenotron and that of the choke and transformer windings. When capacitor C_1 is present, the lowering of the voltage is also attributed to a faster discharge of the capacitor when the load resistance is decreased. On large currents the capacitor gives practically no voltage increase because it discharges too quickly and, as a result, curve *2* almost coincides with curve *1*. Both curves pass through the short-circuit point, where the current is denoted by I_{sc}. Rectifiers, however, are not operated at such heavy currents, because this is harmful to them and uneconomical, as the output voltage is low. Various handbooks give the external characteristics of rectifiers for operation with different types of kenotrons at different capacitance values of C_1 on currents which do not exceed a certain maximum permissible value of $I_{= max}$. The smaller the capacitance, the steeper the curve.

The first filter capacitor (C_1), although it increases the useful output voltage, simultaneously increases the so-called *inverse voltage* across the kenotron, i.e., it increases the negative voltage applied to the valve anode during those half-cycles when the kenotron is not conducting. This can lead to a breakdown of the valve and therefore should be kept within certain limits. In all the rectifier circuits studied above, the maximum value of inverse voltage is obtained on negative peaks of voltage, when the rectifier valve is not conducting.

In half-wave rectifiers the maximum inverse voltage is equal to the sum of the peak transformer voltage E_m and the voltage across the first filter capacitor C_1. The latter of the two voltages is also equal to E_m under no-load conditions, but is much smaller when a load is connected to the rectifier. Hence in a half-wave rectifier the maximum inverse voltage can be assumed to be approximately equal to $2E_m$. Similar values of inverse peak voltage ($2E_m$) are built up across each rectifying unit in the other circuits studied above. In one of these circuits, pertaining to a full-wave rectifier (Fig. 101), the value of E_m is developed by one half of the transformer winding; only in the bridge circuit (Fig. 102), the inverse peak voltage is half as large as in the other circuits, which is an advantage.

If the pulsating voltage change ΔU_1 across the first filter capacitor C_1 does not exceed 10-20% of d.c. voltage $U_{=}$, it can be approximately calculated on the basis of the following considerations. When the capacitor is being discharged and the voltage across it decreases by a certain value ΔU_1, the quantity of electrons supplied by the capacitor to load resistance can be given by $q = C_1 \Delta U_1$. When pulsations are of low level, it can be assumed without great error that the capacitor discharge lasts throughout the whole period of pulsations T_p (actually, the time of discharge is somewhat shorter than a period). The quantity of electricity given by the capacitor creates a pulsating current in which the d.c. component is given by I_{-}. Hence this quantity of electricity can be otherwise represented by $q = I_{-} T_p$. Therefore, $C_1 \Delta U_1 = I_{-} T_p$ or $\Delta U_1 = \frac{I_{-} T_p}{C_1}$.

Substituting $\frac{1}{f_p}$ for T_p, where $\frac{1}{f_p}$ is the frequency of pulsations, we have the following:

$$\Delta U_1 = \frac{I_-}{C_1 f_p}.$$

If, in accordance with Ohm's law, $\frac{U_-}{R}$ is substituted for I_-, where R is the load resistance, we obtain:

$$\Delta U_1 = \frac{U_-}{R C_1 f_p}.$$

As follows from the formula, the greater the load resistance R, the higher the pulse frequency f_p, and the larger the capacitance C_1, the lower will be the value of pulsations ΔU_1. It sould be kept in mind that the above formula can be used for calculations with a fair degree of accuracy only for low pulsations. In practical cases this formula helps to find the value of C_1 securing a given value of pulsations ΔU_1 when the values U_-, R and f_p are known.

Further smoothing of pulsations is provided by the so-called L-type filter, consisting of choke L (or of ohmic resistance R_f) and second capacitor C_2. The value showing by how many times this filter will decrease the pulsations is known as the *smoothing factor* or *filtering factor* k_f. The filtering factor is usually found for the first harmonic of pulsations, but it can be also used for approximate calculation of pulsation value ΔU_2 at the output of the filter. Obviously, ΔU_2 is given by the following:

$$\Delta U_2 = \frac{\Delta U_1}{k_f}.$$

For inductive-capacitive filter (choke-type filter) LC_2 the value of k_f is calculated from the following formula:

$$k_f \approx \omega_p^2 L C_2 = 4\pi^2 f_p^2 L C_2 \approx 40 f_p^2 L C_2.$$

For resistive-capacitive filter $R_f C_2$ the formula is rewritten as follows:

$$k_f \approx \omega_p R_f C_2 = 2\pi f_p R_f C_2 \approx 6.25 f_p R_f C_2.$$

Both formulas, giving the value of k_f, hold true only in those cases when k_f is considerably larger than 1. It so happens that these very cases are of the greatest interest, because a filter whose k_f is only slightly larger than 1 is poor and should not be used.

Example 1. Find the value of voltage pulsation at the output of a filter in which $C_1 = C_2 = 20$ mfd and $L = 40$ h, if the rectifier gives $U_- = 250$ v and supplies power to load resistance $R = 5{,}000$ ohms; $f_p = 100$ cps.

Solution:

$$\Delta U_1 \approx \frac{250}{5\times 10^3 \times 20 \times 10^{-6} \times 100} \approx 25 \text{ v};$$

i.e., the value of voltage pulsation is 10%; therefore the calculation is sufficiently accurate.

The filtering factor of the L-type part of the filter is given by the following:

$$k_f \approx 40 \times 100^2 \times 40 \times 20 \times 10^{-6} = 320.$$

Hence the pulsation voltage ΔU_2 at the output of the filter is found as follows

$$\Delta U_2 = \frac{25}{320} \approx 0.08 \text{ v}.$$

Example 2. If in the example given above the choke is replaced by resistor $R_f = 2{,}000$ ohms, find ΔU_2 for such a case.

Solution:

In this case

$$k_f \approx 6.25 \times 100 \times 2 \times 10^3 \times 20 \times 10^{-6} \approx 25;$$

$$\text{and } \Delta U_2 = \frac{25}{25} = 1 \text{ v.}$$

Obviously the filtering action of a resistive-capacitive filter is far inferior to that of an inductive-capacitive filter.

55. TYPES OF KENOTRON AND THEIR DESIGN

Low-power kenotrons have glass envelopes supplied with bases. The anodes are made of nickel and are shaped as cylinders, and the cathodes are placed inside them. In a twin-anode kenotron both anodes, together with their cathodes, are placed side by side and their heaters are connected in parallel. For better cooling the anodes are frequently supplied with ribs, which increase the anode surface. The latter is blackened to further facilitate the cooling. There are filamentary and indirectly-heated types of kenotrons, the emitting surfaces usually of oxide-coated variety.

Kenotron ratings comprise the following factors:

U_f — filament or heater voltage;
I_f — filament or heater current;
$I_{a\,max}$ — maximum permissible pulse value of anode current;
$I_{=\,max}$ — maximum rectified current;
U_{inv} — peak inverse voltage which the kenotron can stand;
$U_{2\,max}$ — maximum alternating voltage delivered by the secondary (step-up) winding of power transformer;
$P_{a\,dis}$ — maximum power dissipated by the anode.

The first two values need no explanation. The maximum permissible pulse value of anode current ($I_{a\,max}$) is determined by the emission current of the cathode. Maximum rectified current ($I_{=\,max}$) depends upon $I_{a\,max}$, is always smaller than $I_{a\,max}$, and also depends upon the maximum permissible power dissipated by the anode ($P_{a\,dis}$), which prohibits large values of $I_{=\,max}$. Peak inverse voltage U_{inv} is the maximum voltage that can be applied between anode and cathode without setting up an arc between the two electrodes and producing sparking or insulation breakdown.

In all the rectifier circuits studied above, except for the bridge circuit, when capacitor C_1 is provided at the input of a filter, the effective voltage U_2 developed by the secondary winding of transformer must not exceed 35% of U_{inv} value (remember that in full-wave centre-tap rectifiers U_2 is related to one half of the winding). In the case of a bridge circuit U_2 must not exceed 70% of U_{inv}.

Maximum permissible anode dissipation depends upon the design and dimensions of anode, as well as upon the material of which it is made.

Modern Soviet kenotrons employ eight-pin bases. In the marking of kenotrons the first figure indicates a rounded-off value of filament or heater voltage. Twin-anode kenotrons types 5Ц4С and 5Ц4М are widely used. Single anode kenotron type 30Ц1М and twin-anode kenotron type 30Ц6С are both designed for operation in rectifiers without step-up transformers. The last named is provided with separate cathode terminals and can operate in voltage-doubling circuits. Twin-anode kenotrons, types 5Ц3С, 5Ц8С, 5Ц9С, 6Ц5С, 6Ц4П and 6Ц10П are also widely used. Special-purpose single-anode kenotrons, types 1Ц1С, 1Ц7С, 1Ц11П and 2Ц2С are designed for high-voltage rectifier duty (up to several thousand volts) in low-current circuits.

Low-power rectifiers sometimes employ double diodes types 6Х6С and 6Х2П, although these are primarily designed for detector duty in radio receivers. When used as rectifiers they can deliver rectified current not in excess of 10 ma and voltage not greater than 150 v. These valves have two separate cathodes and hence can be used in voltage-doubling circuits.

In a low-power rectifier it is also possible to use an ordinary triode as the rectifier valve. The grid and anode of the valve are connected together in such an application, and the valve is thus converted into a diode. As a precaution a resistor is sometimes connected into the grid circuit in order to limit the power dissipated by the grid and to prevent its overheating.

56. KENOTRON RECTIFIER COMPONENTS

Power transformers are the most important components of kenotron rectifiers. The primary winding of a power transformer usually has several sections which can be switched to operate on different mains voltages (generally 110, 127 and 220 v). The step-up winding is designed to give 250-500 v and sometimes more, depending upon application requirements. Power transformers intended for operation in full-wave rectifiers normally have a centre-tapped secondary winding, with at least two step-down windings in addition to the step-up winding. One of the step-down windings is used for heating the filament or heater of the kenotron, and the other to supply power to the filaments or heaters of other valves of receiver, transmitter or amplifier. Higher-power rectifiers use separate transformers for heating the filaments or heaters of kenotrons and other valves. To suppress the interference reaching the rectifier from the supply mains, a shielding winding is provided in some models be-

tween the primary and secondary windings. One end of the shielding winding is earthed, while the other is free.

A rectifier filter most frequently employs electrolytic capacitors designed for the required working voltage. The cores of filter chokes are usually provided with an air gap. This prevents magnetic saturation, the occurrence of which could seriously lower the inductance. The resistance of choke winding is equal to several hundred ohms. A part of the rectified voltage is lost in this winding, as well as in the step-up winding of the power transformer.

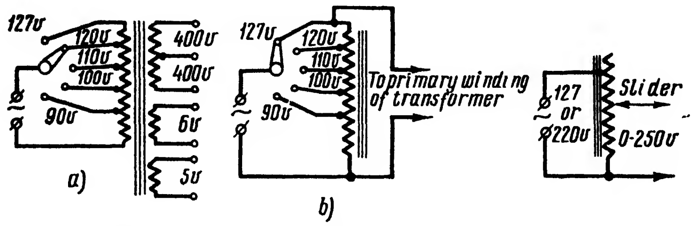


Fig. 111. Methods of voltage adjustment

A switch and a fuse are provided in the primary (mains) winding of the power transformer. The fuse automatically disconnects the transformer from the power mains when a short-circuit occurs. For instance, if a filter capacitor is punctured, thus short-circuiting the rectified-current circuit, the primary winding of the transformer will draw abnormally heavy current from the mains and the fuse will blow. If no fuse were provided, the short-circuit would cause a transformer burn-out.

Sometimes the taps of a sectionalised primary winding are brought out to a step switch (Fig. 111*a*). Setting the switch to various steps, corresponding to different mains voltage (for instance: 90, 100, 110, 120 and 127 v), makes it possible to maintain a constant value of rectified voltage despite fluctuations of the mains. Another way of controlling the output voltage of a rectifier is the inclusion of a special autotransformer between the power transformer and the mains. Such an autotransformer is provided with taps for different mains voltages and, together with a step switch, feeds the correct primary voltage to the power transformer, even though the mains voltage fluctuates (Fig. 111*b*).

The radio industry also manufactures special regulating autotransformers which, when connected to 127 or 220 volt mains, give an output voltage continuously variable between 0 and 250 volts (Fig. 111*c*). In such autotransformers the slider moves directly along the winding turns.

When building or operating a rectifier, be sure to observe safety rules, as several hundred volts can kill. All high-voltage parts of the circuit must be securely protected against accidental contact.

Never touch any parts of an operating rectifier. All experimenting with a rectifier should be done only when the rectifier is disconnected from the mains and its filter capacitors are discharged. A neon lamp is a handy piece of equipment when checking any type of radio equipment for high voltage, since it lights up when connected to a high voltage circuit. It is a good practice to permanently connect such a lamp across filter capacitors (through a limiting resistor of several tens of thousands of ohms). Incidentally, the constant presence of a neon lamp across the high-voltage circuit will constitute a load protecting filter capacitors from breakdown in case of overvoltage. Such overvoltage is liable to take place if for some reason the normal load is disconnected from the rectifier. There will then be no internal voltage drop in the rectifier, because no current will be flowing through the kenotron, step-up winding or power transformer, and filter choke winding; therefore, the voltage across the filter capacitors will rise to its maximum value.

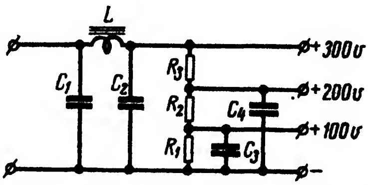


Fig. 112. Method of obtaining several voltages from a single rectifier with the help of a voltage divider

When it is required to obtain several values of rectified voltage from the same rectifier, voltage dividers are employed. Fig. 112 illustrates the connection of such a voltage divider, designed to give three values of voltage. It often happens that lower-voltage circuits (for instance, first amplifier stages) require better filtering than higher-voltage loads. In such a case the illustrated voltage divider will perform an extra function, taking the place of additional filter sections, provided that auxiliary filter capacitors C_3 and C_4 are connected as shown in Fig. 112.

57. FUNDAMENTALS OF POWER TRANSFORMER DESIGN

When a new power transformer for a rectifier is to be built or an old one rewound, simple calculations must first be made. For power transformers of 100-200-watt rating, the design procedure is as follows.

Taking into consideration the voltage and maximum current to be given by the secondary winding of the transformer (U_2 and I_2), we find the secondary circuit power

$$P_2 = U_2 I_2.$$

When the transformer is provided with several secondary windings, P_2 is equal to the sum of powers of all the secondary windings. Assuming a transformer efficiency of 80%, the power of the primary winding is determined

$$P_1 = \frac{P_2}{0.8} = 1.25\ P_2.$$

Power is transferred from the primary to the secondary circuit via the magnetic flux. Hence the cross-sectional area S of a transformer core depends upon the amount of power to be transferred and is proportional to it. When the core is to be made of ordinary transformer steel, S is given by the following formula:

$$S = \sqrt{P_1}.$$

where S is given in square centimetres and P_1 in watts.

From the value of S is determined the number of turns per volt, referring to voltage w'. When transformer steel is used, this number is given by the following equation:

$$w' = \frac{50}{S}.$$

If the transformer core is made of lower-grade materials, e.g., tin, roof steel, steel wire (such materials must be first annealed to make them soft), the values of S and w' should be increased by 20-30%.

The number of turns in all the windings can now be calculated:

$$w_1 = w'U_1; \quad w_2 = w'U_2; \quad w_3 = w'U_3, \text{ etc.}$$

When current flows through a transformer winding, a considerable part of the voltage may be lost because of voltage drop in the winding. Hence it is recommended to wind 5-10% turns more than the calculated value.

The current through the primary winding is as follows:

$$I_1 = \frac{P_1}{U_1}.$$

The diameter of winding wires is determined from the current value; an average current density of 2 amperes per sq. millimetre is usually allowed in small transformers. With such current density the diameter of wire (not considering its insulation) is calculated for any winding, the result being expressed in millimetres:

$$d = 0.8\sqrt{I}.$$

If no wire of required diameter is available, the winding can be made with several thinner wires, connected in parallel. The total cross-sectional area of such wires must not be less than the computed area of the required wire. The cross-sectional area of a wire is given by the following equation:

$$q = 0.8d^2.$$

Filament windings are usually made with a few turns of thick wire and are placed on top of all the other windings. The current density for such filament windings can be increased up to 2.5 and even 3 a/mm², because these windings are subjected to better cooling. In this case a coefficient of 0.7 or 0.65 should be substituted for 0.8 in the formula used for finding the wire diameter.

In conclusion, check the location of windings in the space provided for them. The total cross-sectional area made up by the turns of each winding can be found by multiplying the number of turns by the cross-sectional area of the wire. The latter is taken as $0.8\,d_{ins}^2$, where d_{ins} stands for diameter of wire, allowance having been made for the insulation. The cross-sectional areas of all the windings are added up. Then the total area thus determined is increased two or three times in order to make an approximate allowance for such factors as loose parts of windings, coil forms, insulating spacers between the windings and between separate sections of windings. The area of core openings must not be less than the value arrived at by calculation.

An example of power transformer design. Using the above procedure, let us design a power transformer for a rectifier used for supplying a radio receiver. Assume that the transformer will have a high-voltage winding consisting of

two 300-volt 50-milliampere sections, and that the same transformer is to have two low-voltage windings, one of which is rated at 5 volts and 2 amperes, and the other at 6.3 volts and 1.35 amperes. The mains voltage is taken equal to 220 volts.

First we determine the total power of all the secondary windings:

$$P_2 = 2\times 300\times 0.05 + 5\times 2 + 6.3\times 1.35 = 30 + 10 + 8.5 = 48{,}5 \text{ watts.}$$

The power of the primary circuit is as follows:

$$P_1 = 1.25\times 48.5 = 60 \text{ watts.}$$

The cross-sectional area of the core, made of transformer steel, is found next:

$$S = \sqrt{60} \approx 7.7 \text{ sq. cm.}$$

To determine the number of turns per volt:

$$w' = \frac{50}{7.7} = 6.5.$$

Current drawn by the primary winding:

$$I_1 = \frac{60}{220} = 0.27 \text{ amperes.}$$

The number of turns and diameter of wires in each winding are as follows: in the primary winding—$w_1 = 6.5\times 220 = 1430$; $d_1 = 0.8\sqrt{0.27} = 0.41$ mm; in the step-up winding—$w_2 = 6.5\times 2\times 300 = 3{,}900$ (round off this figure to 4,000); $d_2 = 0.8\sqrt{0.05} = 0.18$ mm; in the kenotron filament winding—$w_3 = 6.5\times 5 = 32.5$ (round off to 35); $d_3 = 0.65\sqrt{2} = 0.93$ mm; in the winding designed to supply receiver heaters—$w_4 = 6.5\times 6.3\times 6.3 = 41$ (round off to 45); $d_4 = 0.65\sqrt{1.35} = 0.74$ mm.

Now assume that the core opening designed to accommodate the windings has a cross-sectional area of $6\times 3 = 18$ sq. cm or 1,800 mm², while the diameters of insulated wires selected for the windings are as follows:

$$d_{1ins} = 0.44 \text{ mm};\ d_{2ins} = 0.2 \text{ mm};\ d_{3ins} = 0.98 \text{ mm};\ d_{4ins} = 0.8 \text{ mm}.$$

Now let us check the location of windings in the core opening. First we find the cross-sectional area of the windings:

primary winding: $0.8\times 0.44^2\times 1{,}430 = 250$ mm²;
step-up winding: $0.8\times 0.2^2\times 4{,}000 = 128$ mm²;
kenotron filament winding: $0.8\times 0.98^2\times 35 = 27$ mm²;
receiver valves filament winding: $0.8\times 0.8^2\times 45 = 23$ mm².

The total cross-sectional area taken up by all the windings is about 428 mm². Obviously this is over four times smaller than the core opening; consequently, all the windings will fit in.

58. GAS-FILLED VALVE RECTIFIERS, THYRATRONS AND IONIC VOLTAGE STABILISERS

A gas-filled valve rectifier is an ionic device with so-called arc discharge. This rectifier is represented by a diode containing an incandescent activated cathode. Mercury vapour or some inactive gas is contained in the rectifier envelope at a pressure of 0.01-0.001 mm of mercury column. Let us first examine the action of a gas-

filled valve rectifier with mercury vapour, otherwise known as a mercury-vapour rectifier.

As in all ionic devices, ionisation takes place also in a mercury-vapour rectifier; flying electrons knock out new electrons from the neutral atoms of mercury vapour, the atoms being converted into positive ions. Consequently, a mercury-vapour rectifier possesses several advantages over a kenotron. The anode current is many times greater because of the obtaining of additional free electrons by ionisation, and because of the neutralisation of the negative space charge by positive ions. The internal resistance of a normally operating mercury-vapour rectifier is extremely low and may be equal to several ohms. Voltage drop in this type of rectifier is constant and equals only 9-12 volts.

If the initial anode voltage applied to a mercury-vapour rectifier is gradually increased from zero to some small value, the anode current will be at first very low, because the ionisation has not yet begun and the process taking place in the rectifier thus far is purely electronic. Ionisation suddenly appears when the anode voltage reaches approximately 10 volts. This critical voltage is known as *the firing voltage* of a mercury-vapour rectifier and is characterised by the appearance of bluish-violet glow in the rectifier and by a sharp increase of its anode current. A further increase of the supply voltage does not increase the anode voltage across the valve but leads to an increase of current in the circuit, resulting in a greater voltage drop across the load resistor. It should be remembered that such a load resistor should be always kept connected in series with a mercury-vapour rectifier. If the resistor is omitted, full voltage of the anode power supply will be applied to the mercury-vapour rectifier. Because of extremely low resistance of the rectifier, its anode current will immediately increase to such a high and quite impermissible value that the effect can be likened to a short-circuit and the rectifier will be damaged.

A mercury-vapour rectifier is noted for the following disadvantages. Its cathode will be destroyed if its heating temperature is below the rated value, as will be understood from the following. At lower temperatures the cathode emission decreases and the ionisation becomes weaker, so that the electron space charge is not fully neutralised. The internal resistance of the rectifier valve increases, as does the voltage drop between anode and cathode, and positively-charged ions begin to travel at higher velocities. Consequently, the cathode bombardment by the ions is intensified to the point where the ions destroy the active layer of the cathode. Accordingly, the filament of a mercury-vapour rectifier should never be run at a voltage reduced by more than 5% from the rated filament voltage value. If the rectifier service life is not to be shortened noticeably, this voltage also should not be allowed to exceed the rated value by more than 10%.

Normal mercury-vapour pressure in a rectifier of this type is obtained only when the ambient temperature is between 15 and 50°C. The filament and anode supplies of a mercury-vapour rectifier should never be switched on at the same time. It is absolutely necessary that the filament supply is switched on first and the rectifier is properly heated up for about a minute or more (depending upon its type); only after this may the anode supply be switched on. If this is not observed, the following will happen. When a mercury-vapour rectifier is not sufficiently heated up, the gas pressure in it is below the required value: a lowered pressure causes only weak ionisation; the anode resistance of the valve remains high, while the voltage drop inside it is considerable. This is the same type of situation as when a mercury-vapour rectifier operates on abnormally low filament voltage; the ions, travelling with excessive velocity, bombard the cathode with sufficient intensity to destroy it.

It should be easy to understand from the foregoing why the anode voltage should be switched off before the filament voltage when it is required to shut down a radio set employing mercury-vapour rectifiers.

Here is the explanation of another "tender point" of mercury-vapour rectifier operation. The ionic process taking place within this type of rectifier offers a possibility of so-called inverse current action. When the rectifier is in service and its anode voltage drops to zero (which happens twice during every alternating-current cycle), the ions are not converted to neutral molecules instantaneously, because such deionisation requires from 0.0001 to 0.00001 of a second. As the anode becomes increasingly negative, the ions begin to move in a reverse direction, travelling towards the negatively-charged anode and thus forming an inverse current. If the voltage is sufficiently high, such inverse current can cause "inverse ignition", i.e., an arc discharge. This interferes with the rectification process taking place in the valve. The maximum value of permissible inverse voltage is specified for each type of mercury-vapour rectifier. It should be noted that presence of mercury drops on the anode of the valve provokes the inverse ignition effect, because such drops can become sources of emission. To prevent this all new mercury-vapour rectifiers should be thoroughly heated for a considerable length of time when first put into service. This is done by switching on the rectifier filament (anode voltage must be off) for half an hour or more, until no drops of mercury are visible in the valve, all of the mercury having turned into vapour. Only after this may the anode voltage be applied to the rectifier.

Despite their shortcomings, mercury-vapour rectifiers are widely used because they offer advantages not possessed by kenotron rectifiers. The mercury-vapour rectifier is but one of several varieties of gas-filled valve rectifiers whose design features are summed up below.

Mercury-vapour rectifiers. Fig. 113*a* shows the general construction principle of rectifiers of this type. The same figure also gives a schematic representation of gas-filled valve rectifiers; the shading of a dot indicates that the valve envelope contains gas.

Mercury-vapour rectifiers of comparatively low power often employ filamentary oxide-coated emitters. In order to prevent an arc between filament ends, the filament voltage is made low and does not exceed a few volts. Small mercury-vapour rectifiers are provided with bases similar to those used by lighting bulbs or by usual radio valves.

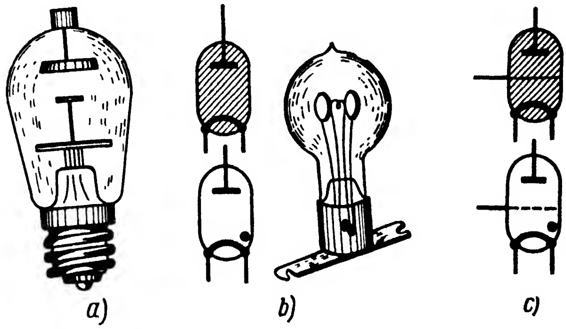


Fig. 113. External view and schematic representation of mercury vapour rectifiers. Schematic representation of thyratrons

The anodes of low-power rectifiers are made of nickel and are disc-shaped. Semi-spheric graphite anodes are employed by higher-power rectifiers. As a rule, the anode terminal is brought out through the top of the valve envelope. The lower part of the envelope usually contains several drops of mercury. Since the pressure of mercury vapour must not be too high, a disc screen is provided in the valve to decrease mercury heating. Another metal screen (shaped as a disc or a cylinder) is also placed in the valve envelope for the following purposes: 1) to decrease the thermal losses of the cathode, 2) to protect the cathode from bombardment by positively-charged ions, 3) to preclude the possibility of reaching the anode by chips off the cathode active layer, thus eliminating a serious cause for inverse ignition.

Mercury-vapour rectifiers are nearly always manufactured in single-anode variety and are designed for application in practically any type of rectifying circuits. They have found a particularly wide application in high-voltage rectification systems.

A recent development in the field of gas-filled valve rectifiers has pointed a way to filling the envelopes of such rectifiers with some inert gas, instead of mercury vapours. Xenon and krypton

have proven to be the most suitable gases in such application. When the rectifier envelope is filled with such inert gases, the firing voltage, depending upon the type of gas, is increased to a value between 15 and 25 volts (as compared to 10 volts when the envelope is filled with mercury vapour). The voltage drop value across an operating rectifier of such type is approximately equal to the same figure.

Argon-filled valve rectifiers. Rectifiers of this type are employed for full-wave rectification of low voltages. They employ filamentary emitters made of thoriated molybdenum, and their anodes are made of nickel. The external view of such rectifiers is shown in Fig. 113*b*. In comparison with mercury-vapour rectifiers, argon-filled rectifiers have the advantage of permissible operation at low and high ambient temperatures (from —30 to +50° C). Moreover, their heating-up time is shorter (about 20-30 seconds). In some circuits simultaneous application of filament and anode voltage to argon-filled valves is permissible. Rectifiers of this type are used for battery-charging, supplying power to sound amplifiers, etc.

All types of gas-discharge valve rectifiers must be provided with separate filament and anode switches. Smoothing filters employed with these amplifiers must begin with a choke (choke-input filters) and not with a capacitor. Capacitor-input filters are not permissible in this case because, owing to the extremely low internal resistance of the rectifier, the pulses of anode current charging the input capacitor would be excessively high and could damage the gas-discharge valve rectifier.

Thyratrons. A thyratron is a gas-discharge valve rectifier supplied with a grid. This may be called a gas-discharge triode, and its schematic representation is shown in Fig. 113*c*.

A thyratron differs from an ordinary triode valve in one interesting respect; the thyratron grid cannot control the anode current, which, once started, will continue to flow no matter how high the negative voltage applied to the grid.

However, changing the value of negative potential on the grid of a thyratron offers a method of controlling the firing. The higher the negative voltage of the thyratron grid, the higher will be the anode voltage required to ignite the valve.

As noted above, the grid will have no control over the anode current, for when the anode current begins to flow (after the valve has fired), the negatively-charged grid attracts positive ions, and the latter surround the grid and neutralise its action.

The cathode and anode of a thyratron are similar to those of other gas-discharge valves. The thyratron grid completely surrounds the cathode, and all the electrons moving from the cathode to anode have to pass through it. Low-power thyratrons are quite similar to ordinary vacuum triodes as far as the electrode design is concerned.

Depending upon the type, the normal operating voltage of a thyratron is between 10 and 30 volts. This voltage can be much higher before the firing of the thyratron (if the grid has been maintained at a high negative potential).

The anode current of a thyratron which has fired can be stopped by interrupting the external circuit feeding anode voltage to the valve. The anode current can be also stopped by reducing the anode voltage to its extinguishing value, which is only slightly lower than the normal anode voltage of an operating thyratron.

Thyratrons have found application in controlled rectifying circuits, where by changing their grid potential it is possible to regulate rectified voltage from zero up to maximum without any additional energy losses. Thyratrons are also used as relays (where the frequency of controlled circuits does not exceed several tens of kilocycles). Numerous other applications are constantly being found for these gas-discharge triodes.

Among the various gas-discharge valve rectifiers the most popular types are: ВГ-161 and ВГ-129 (single-anode mercury-vapour valves); ГГ1-0.5/5 (single-anode valve with xenon-krypton gas mixture); ВГ-176 (twin-anode argon-filled valve); ГР1-0.25/1.5 (twin-anode mercury-vapour valve).

Among low-power thyratrons filled with inert gases the most popular types are: ТГ-212; ТГ-213; ТГ1-0.1/0.3; ТГ1Б; ТГ1-0.1/1.3; ТГ3-0.1/1.3.

Besides these, radio manufacturing industry produces many different types of high-power gas-discharge valve rectifiers and thyratrons.

According to the present system of marking in the U.S.S.R., the marking formula begins with letter Г in case of gas-discharge valve rectifiers, and with letter Т in case of thyratrons. The next letter shows whether the valve is filled with mercury vapour (Р) or with inert gases (Г). Then comes a number helping to distinguish various types of ionic devices for which the other symbols of the marking formula are similar. Given at the end of the formula is the maximum rectified current (d. c.) in amperes, and, after the solidus, peak inverse voltage in kilovolts.

For instance, formula ТГ1-0.1/0.3 stands for the following specifications: thyratron, filled with inert gas, number one, maximum rectified current — 0.1 a, peak inverse voltage — 0.3 kilovolt.

Ionic voltage stabilisers. Ionic stabilisers are employed when it is required to maintain a constant value of rectified voltage. These utilise the glow discharge phenomenon and are represented constructionally by two cylinders, one of which is inserted into the other. These are placed in an envelope containing low-pressure argon or neon. The electrode with the greater surface serves as a cathode. When properly operated such stabilisers can maintain a constancy of voltage within 1-3 volts. A limiting resistor R is

connected in series with a stabiliser (Fig. 114). The curve in Fig. 114*b* shows the relation between output voltage U_0 of the stabiliser and input voltage U_i. When U_i is made to increase, U_0 also increases. The stabiliser fires at a certain voltage. When this occurs U_0 is somewhat reduced, because current begins to flow through the stabiliser and additional voltage drop takes place across resistor R. A glow appears at some part of the cathode. Within certain limits a further increase of U_i produces practically no change of U_0, the latter remaining constant and approximately equal to the working voltage U_w of the stabiliser.

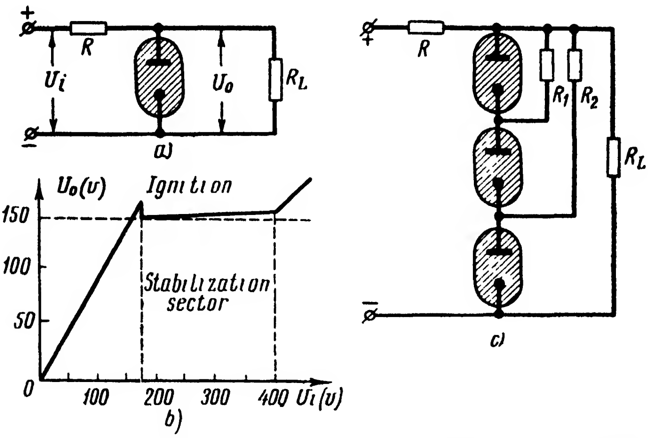


Fig. 114. The circuit of ionic voltage stabilisers and stabilisation characteristic

As U_i is made to increase further, the current through the stabiliser also increases and the glow continues to spread over greater part of the cathode surface. It is significant, however, that the brightness of the glow does not change. This is the condition under which voltage stabilisation takes place, such a condition being called the *condition of normal cathode drop* (here "drop" pertains to potential drop near the cathode).

A simplified explanation of voltage stabilisation is as follows. As the current increases, the operating (current-carrying) area of the cathode is also increased. This correspondingly reduces the internal resistance of the stabiliser. The voltage drop across the stabiliser (current times resistance) remains practically unchanging.

If U_i is made to increase still further, there will come a moment when the entire cathode surface will begin to glow. Under such a condition, which is known as the *condition of abnormal cathode drop*, voltage stabilisation no longer takes place; current increase will now cause voltage increase.

A voltage stabiliser of the described type has the following parameters:

U_w — working voltage;
I_{min} — minimum current;
I_{max} — maximum current (which, together with I_{min}, determines the limits of stabilised condition);
U_f — firing voltage (which is usually from 5 to 25 volts higher than U_w).

A correct stabilised condition is secured only when R has a definite value of resistance, calculated from the following formula:

$$R = \frac{U_{av} - U_w}{I_{av} + I_L},$$

where:

U_{av} — average value of input voltage;
U_w — working voltage of stabiliser;
I_{av} — average value of stabiliser current (this value is equal to $\frac{I_{min} + I_{max}}{2}$);
I_L — load current (this value is equal to $\frac{U_w}{R_L}$).

The higher the value of resistance R, the broader will be the limits of supply voltage change over which stabilisation will take place. From the above-given formula for R it is clear that large values of R are obtained on higher voltage of the supply and on lower load currents.

Soviet industry produces voltage stabilisers with eight-pin bases. These are rated at $I_{min} = 5$ ma and $I_{max} = 30$ ma, and are designed for working voltages of 75, 105 and 150 v. The types are: СГ2С, СГ3С, СГ4С. Besides these, our industry also produces bantam stabilisers designed for 150 v operation (type СГ1П), 105-v bantam stabilisers (СГ2П), and miniature 150-v stabilisers (СГ5Б).

For operation on high voltages stabilisers are connected in series (Fig. 114*c*). In such circuits shunt resistors R_1 and R_2 (0.5-1 megohm) are connected across stabilisers to facilitate their firing. Stabilisers are never connected in parallel, because various stabilisers of the same type have slightly different values of firing voltage; if two were connected in parallel, the one with lower firing voltage would fire first and would not allow the other stabiliser to operate.

It is interesting to note that ionic voltage stabilisers offer negligible internal resistance to alternating current and are very effective in smoothing out the pulsations of rectified currents. In this respect they serve just as well as filter capacitors with very high value of capacitance.

Energy losses in the stabilisers and associated resistors R represent a disadvantage of ionic stabilisers. Therefore they are not manufactured in high ratings and are used only to stabilise comparatively low-power equipment in which high efficiency is not of paramount importance.

59. SEMICONDUCTOR RECTIFIERS

Cuprous-oxide rectifiers consist of separate rectifying units shaped as round or rectangular copper plates, one side of each being covered with cuprous oxide. The plate is one terminal of a unit, the other being a lead plate bearing against the cuprous-oxide layer (Fig. 115*a*).

Cuprous oxide is a semiconductor. At the point where it contacts the copper a so-called *barrier-layer* is formed. The resistance of

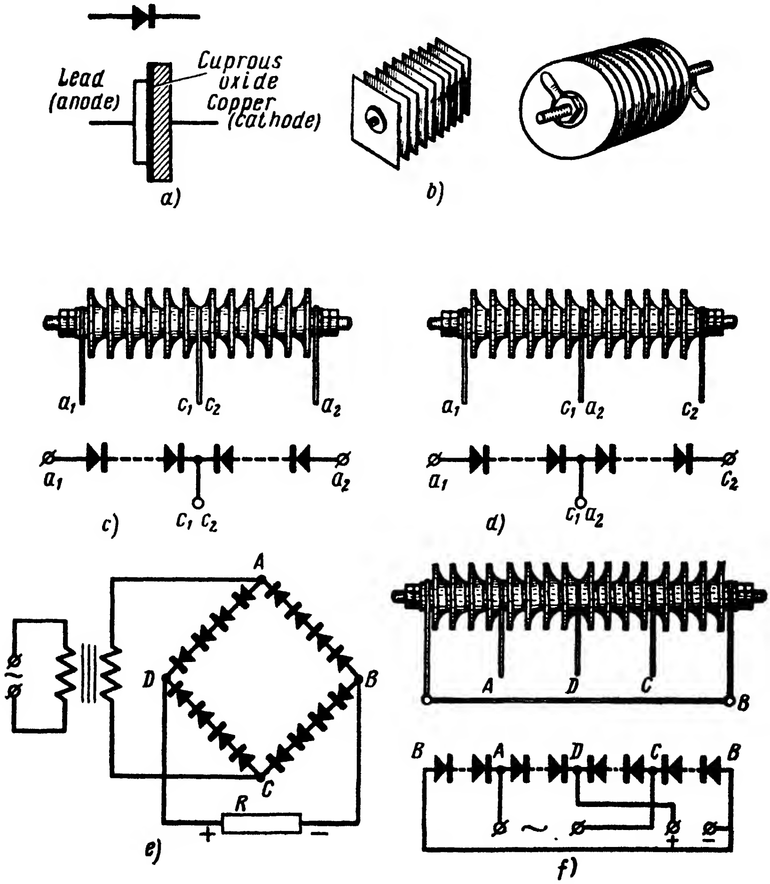


Fig. 115. Construction of cuprous-oxide and selenium stacks, and a bridge rectifier circuit employing such rectifiers

the barrier layer is low when the copper is negative in respect to the cuprous oxide. However, if the polarity is reversed this resistance is increased hundreds of times. In the unit being described the copper plate is the cathode, while the cuprous-oxide layer acts as the anode.

When a cuprous-oxide unit is used as a rectifier, the alternating voltage value across it should not exceed 12 volts, otherwise the barrier layer will be punctured. This is the main disadvantage of cuprous-oxide rectifiers. When they are used on higher voltages, a corresponding number of cuprous-oxide units have to be connected in series. Rectifiers of this type should not be subjected to temperatures in excess of 50° C, or their inverse current will become excessive and a breakdown of the rectifier can result. The permissible current density is 50 ma to each sq. cm of the operating surface of the copper plate.

Constructionally the units of a cuprous-oxide rectifier are assembled into bolted stacks. When better cooling of such stacks is required, larger radiator plates are sandwiched between the units (Fig. 115*b*). In respect to the load, the copper plate of a cuprous-oxide rectifier represents the positive terminal of rectified voltage, and the cuprous-oxide layer is then the negative terminal.

With proper maintenance a cuprous-oxide rectifier will give long-term service. It has high efficiency, is reliable in action and convenient in operation. Its properties are impaired, however, by heating and time.

Selenium rectifiers. In comparison with cuprous-oxide rectifiers, selenium rectifiers give better performance and have recently almost completely replaced the former.

A selenium rectifying element is a steel or aluminium plate, one side of which is covered with a layer of semiconductor material—selenium. The selenium surface is coated with fusible metal alloy. On the boundary between this alloy coat and the selenium, a rectifying barrier layer is formed. The metal on the selenium surface becomes a cathode (positive terminal, as far as the rectified voltage is concerned), while the steel or aluminium plate becomes an anode (negative terminal in respect to the rectified voltage).

In selenium rectifiers maximum current density is about 30 ma per 1 cm^2, while the highest permissible value of inverse voltage is 22-25 v for each rectifying unit. Because of such ratings, selenium rectifier stacks contain one-half the number of elements, as compared to cuprous-oxide rectifiers designed for similar voltage. Selenium rectifiers can operate at temperatures as high as 70 ° C. Their service life is measured in several thousands of hours. Compared with cuprous-oxide rectifiers, they have smaller dimensions and weight but higher efficiency. Selenium rectifier stacks with various number of units are assembled for different applications and voltages.

Selenium rectifying elements now manufactured range in diameter from 5 to 130 mm and are designed, respectively, to handle rectified current values from 2 to 3,000 ma. Square and rectangular-shaped units are also used.

The resistance of selenium units is low, e. g., the most frequently used elements with diameters of 18, 25 and 35 mm have respective resistance values of 35, 10 and 4 ohms on maximum permissible values of current. These values slightly increase on smaller currents.

Selenium stacks may be used in all types of rectifying circuits. Normally the units of a selenium stack are assembled for connection to a definite type of rectifying circuit; thus Fig. 115*c* shows a stack designed for operation in a full-wave rectifier circuit, while Fig. 115*d* illustrates a stack to be used in a voltage-doubling rectifier.

Selenium stacks are also often arranged for operation in bridge rectifier circuits. In this case if it is required, for instance, to rectify a voltage of 15-18 v, four units will suffice. On higher voltages each arm of the bridge circuit requires an appropriate number of series-connected units (Fig. 115*e*). The type of stack illustrated in Fig. 115*f* should be used in this case.

The parameters of selenium rectifiers are impaired by time and heating. It is customary to say that they "become aged" when, with the passing of time, their resistance is gradually increased, which results in a greater internal voltage drop and in reduced voltage output of the rectifier.

Recently other types of semiconductor rectifiers — germanium and silicon rectifiers — have come to the fore. They will be discussed in greater detail in Chapter XI.

60. VIBRAPACKS

In some cases radio equipment is supplied with power from a low-voltage storage battery which gives only 6 or 12 volts. This is the usual situation in automobile radio receivers, mobile radio stations, village radio sets with wind-charges, etc.

Such a voltage is sufficient to supply the filaments and heaters of the radio equipment in such installations. But where may the high voltage for the anode circuits be obtained? In such cases *vibrapacks*, described below, usually provide the answer.

A vibrapack is an electro-mechanical device designed for the conversion of d. c. low voltage into d. c. high voltage. Its main component is a vibrator (sometimes called interruptor). Besides the vibrator, every vibrapack also has a step-up transformer, some type of rectifying facility, a smoothing filter and a filter designed for the suppression of interference caused by sparking in the vibrator.

As an illustration, Fig. 116 shows two typical vibrapack circuits. Let us first examine the operation of the low-voltage side of such circuits, similar for both types of vibrapacks.

When switch S is off vibrator armature A, supported on a spring, is in its middle position. But the storage battery current will now flow through one half of transformer primary winding and through the coil of electromagnet EM. Although this current is rather low (electromagnetic coil has high resistance), it is sufficient to make the electromagnet attract armature A and the latter will touch contact C_1. This will short-circuit the electromagnet coil and cause a sharp increase of current flowing through one half of transformer primary winding. The electromagnet, its coil shorted, will release the armature, and the latter, being returned to its original position by the spring, will pass this position because of inertia and will touch contact C_2. This will close another circuit, and the storage battery current will flow through the other half of transformer primary winding.

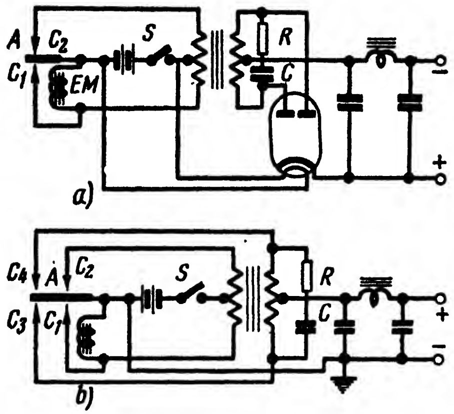


Fig. 116. Vibrapack circuits

However, by this time current also flows through the electromagnet coil, and the electromagnet again attracts the armature. The armature moves away from contact C_2, again touches contact C_1, short-circuits the electromagnet coil, and the above-described process is repeated.

This process keeps on repeating without a stop, as long as switch S is closed. As a result the armature is maintained in a state of continual vibration (with a frequency of 60-200 times per second), alternately and regularly touching contacts C_1 and C_2. Such operation of the armature sends current pulses alternately through the two halves of the transformer primary winding, and this generates a stepped-up alternating voltage in the secondary winding of the transformer.*

High alternating voltage generated by the secondary winding of the transformer has to be rectified into direct voltage before it can be used by radio equipment. In a vibrapack such rectification is generally performed in one of the following two ways:

* The above description gives a very simplified description of vibrator operation. The processes actually taking place in the vibrator are of more complex character.

Fig. 116*a* illustrates how the alternating voltage is rectified by means of a twin-anode kenotron, such as type 6Ц5С, in which the indirectly heated cathode is insulated from the heater.

Fig. 116*b* offers an interesting alternative—an electromechanical rectification. Two additional contacts C_3 and C_4 are provided in this circuit. These are connected to the ends of the secondary winding and play the role of anodes of a kenotron valve. The centre point of the secondary winding serves as the positive terminal of rectified voltage. The negative rectified voltage is obtained from that secondary winding end which at the moment is being contacted by the armature (through either contact C_3 or C_4).

An *RC* circuit is connected across the secondary winding for the purpose of suppressing voltage surges and diminishing sparking at the contacts. The same purpose is served by capacitors connected across each half of the primary and secondary windings.

Despite these measures, sparking still takes place at the contacts and creates considerable interference to radio reception. This necessitates taking additional interference suppression measures such as connection of filters, consisting of high-frequency chokes and capacitors, into the wires carrying rectified current and also into the wires running from the vibrapack to the storage battery. These filters are not shown in Fig. 116, but they are frequently used and their specifications are given in Chapter IX in the section dealing with interference suppression. The vibration unit itself is usually placed into a shielded case, and the whole vibrapack system is thoroughly screened.

Short service life (about 1,000 hours) is the big disadvantage of a vibrapack. At the end of such term the contacts are worn out, and the vibrator unit then has to be replaced by a new one.

61. CURRENT STABILISERS (BARRETTERS)

Special current stabilisers, known as barretters, are sometimes employed in radio equipment for the purpose of keeping the filament current of electron valves at a constant value.

A barretter consists of steel wire placed in a hydrogen-containing envelope. When a voltage increase occurs within certain limits, the resistance of the barretter is also increased in such a way that the current flowing through it remains nearly constant. In practical barretter-stabilised circuits, a change in voltage by as much as 2 times changes the current by less than 5%. Fig. 117 gives a schematic representation of such stabilisers and shows their characteristic.

In barretter marking formulae the first figure gives the current value in amperes. Then comes letter Б (meaning barretter). The formulae end in two figures, written through a hyphen; the first

gives the minimum voltage across barretter terminals, and the second figure the maximum value of such voltage.

The following types of barretter are used at present: 1Б5-9; 1Б10-17; 0,3Б17-35; 0,3Б65-135; 0,425Б5.5-12; 0,85Б5.5-12; 0,24Б12-18.

0.3-ampere barretters are designed primarily for radio receivers using no power transformers. In this application the heaters of all receiver valves are connected in series with each other and with a dropping resistor, the value of this resistor being calculated in the following example.

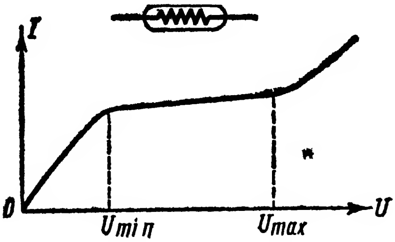


Fig. 117. Current stabiliser characteristic and its schematic representation

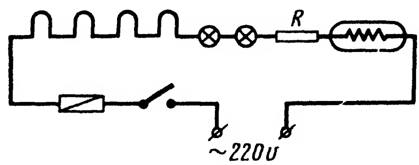


Fig. 118. Radio receiver heater circuit supplied with power from the mains through current stabiliser

Let us assume that we are given a radio receiver in which a selenium rectifier is used instead of a kenotron one. The following electron valves are employed in the receiver, the heaters of all these valves designed to operate on 0.3 a: 6A7, 6K7, 6Г7 and 30П1С. The receiver also employs two scale-illuminating bulbs rated at 6 v and 0.3 a each. All these valves and bulbs are connected in series and are intended for operation from 220-volt mains through type 0.3Б65-135 barretter (Fig. 118).

The total voltage drop across the bulb filaments and across the heaters of all the valves of the receiver is given by: $3 \times 6.3 + 30 + 2 \times \times 6 \approx 61$ v.

Considering 220 volts as the maximum voltage of the mains, and assuming that on this voltage the voltage across the barretter will be equal to 135 volts, we apply Ohm's law to find the value of the dropping resistor in the following way:

$$R = \frac{220 - 135 - 61}{0.3} = \frac{25}{0.3} = 80 \text{ ohms.}$$

If the mains voltage is lowered by 70 v, i.e., down to 150 v, the barretter will still maintain a current of 0.3 a and the operating condition of valve heaters and bulb filaments will remain normal.

It should be noted that barretters possess considerable thermal inertia and are not suitable for stabilisation in circuits where rapid voltage changes take place.

62. QUESTIONS AND PROBLEMS

1. What are the advantages and disadvantages of the usual full-wave rectifier circuit in comparison with the circuit of a half-wave rectifier?

2. Draw a circuit of a full-wave rectifier employing a twin-anode cathode-type kenotron and a filter consisting of two sections.

3. Draw a bridge rectifier circuit with two twin-anode kenotrons, each one of which has two separate indirectly heated cathodes.

4. Devise a circuit for a half-wave rectifier using no power transformer. In this circuit, in series with the kenotron heater connect the heaters of four receiver valves, supplied with power from the given rectifier. Include a filter in the rectified current circuit.

5. The filter of a full-wave rectifier employs an 8-mfd capacitor and a 40-h choke. Find the reactance presented by the capacitor and choke to the a.c. component of the rectified current on 50 cps.

6. The step-up secondary winding of a transformer feeding a half-wave rectifier delivers a voltage of 360 v. What will be the voltage across the filter capacitors if the load resistance is disconnected from the rectifier?

7. Why does the output voltage of a rectifier decrease with the decrease of rectifier load resistance (i.e., with the increase of current flowing through the load)?

8. In what cases can a resistor be substituted for the filter choke in rectifier filters?

9. The load is disconnected from a half-wave rectifier supplied with a filter. The secondary winding of transformer used with this rectifier circuit delivers 280 volts. What will be the value of inverse voltage in such a case?

10. Which part of a rectifier can be damaged if the first filter capacitor (the input capacitor) is punctured?

11. Calculate the limiting resistor for ionic voltage stabiliser type CГ4C if the supply voltage fluctuates between 180 and 220 volts and the load current is 7.5 ma.

12. In what cases are radio receivers supplied with power from vibrapacks?

13. Why are kenotron rectifiers not used for battery charging?

14. Is the fourth curve given in Fig. 101*d* representative of rectified current i_R at the output of voltage-doubling rectifier?

15. The maximum value of the output current of a half-wave rectifier without filter is 20 ma. Find the value of the d.c. component and the amplitude of the a.c. component.

16. Can a moving-iron voltmeter be used for measuring d.c. voltage component across the first filter capacitor?

17. It is required to build a rectifier giving 1,000 volts at 2 milliamperes. Which circuit should be used: half-wave or full-wave?

18. Is a short in the rectified current circuit more dangerous to a kenotron rectifier or to a gas-discharge rectifier?

19. The step-up winding of a rectifier transformer develops 300 volts. The rectified current value is 50 milliamperes. Is it possible to calculate the power dissipated by the anode of the rectifier kenotron as a product of 300 v and 0.05 a (300×0.05)?

20. Why do gas-discharge rectifiers (mercury-vapour valve, thyratron and others) have low internal resistance?

21. Draw the external characteristics of a semiconductor rectifier when operated without filter and when using a capacitor as a filter.

22. Why is it impossible to cut off the anode current of an operating thyratron by the simple expedient of negative grid voltage increase?

23. Design a step-down transformer to feed amplifier valve filaments if the primary voltage is 220 v, and the voltage and current in the secondary circuit are, respectively, 6.3 v and 4 a.

24. What must be the capacitance value of a capacitor shunting a 20,000-ohm load resistor of a half-wave rectifier so that on rectified voltage of 500 v

the voltage of pulsations does not exceed 25 v when the rectifier is operated from 50 cps mains?

25. A full-wave rectifier is supplied with the type of filter shown in Fig. 107*a*. The filter employs 10-mfd capacitors and a 50-h choke. The rectifier delivers 300 volts at 60 milliamperes and operates from 50 cps mains. Find the value of pulsation voltage at the output of the rectifier.

26. Draw a circuit diagram of a rectifier with 5-fold voltage multiplication and with an additional filter designed for smoothing the pulsations.

27. Why does the inverse voltage across a kenotron increase nearly twice when the smoothing filter begins with capacitor? In what type of rectifier circuit such an increase of inverse voltage does not take place?

CHAPTER VI

ELECTROACOUSTIC DEVICES

63. PROPERTIES OF SOUND. THE SENSE OF HEARING

Acoustics is the science of sound. Electroacoustics is a branch of engineering concerned with the conversion of sound oscillations into electrical oscillations and vice versa. It also embraces sound recording and the reproduction of sound with the help of various electrical devices.

In radio engineering an important role is played by different electroacoustic equipment, such as earphones, loudspeakers, microphones, gramophone pickups, etc. Our study of these devices will begin after we have first examined the properties of sound and the peculiarities of the ear.

A source of sound is always represented by a vibrating body, e.g., by a string of a musical instrument. *A sound wave propagated from a source of sound is the vibratory movement of the particles of a resilient medium (e.g., the air) through which the sound travels.* The vibrations taking place in such a resilient medium are conveyed by some of its particles to others and are propagated in all directions from the source of sound. The speed of sound in the air is about 340 metres per second. It is higher in liquids and still higher in the solids.

All sounds are characterised by their *frequency* and *amplitude*, which are their main indices. They are also classified in respect to their pitch and loudness. Examples of high-pitched sounds are: a high-toned whistle, squeaks, a woman's voice (soprano); examples of low-pitched sounds are: a man's voice (bass), drum beats, etc. *The higher the frequency of a sound, the higher is its pitch and vice versa. The larger the sound amplitude, the stronger and louder is the sound.*

It is customary to classify sounds as simple and complex. *A simple sound is noted for the sinusoidal shape of its oscillations.* Sounds possessing oscillation shape other than sinusoidal are complex sounds.

*Every complex oscillation actually represents a sum of several simple oscillations, which are called harmonics and have different amplitudes and frequencies.**

* When referring to sound oscillations, the harmonics are sometimes referred to as *overtones*.

Harmonic frequencies are the multiples of a complex oscillation frequency. For instance, if a complex oscillation has a frequency of 200 cps, its *first harmonic or the basic oscillation* also has a frequency of 200 cps, the *second harmonic* frequency is twice as high — 400 cps, the *third harmonic* frequency is three times as high — 600 cps, etc.

The amplitudes of various harmonics vary and are not governed by such a systematic law as the frequencies. Higher harmonics have smaller amplitudes, but there can sometimes be exceptions. It happens frequently that certain harmonics are totally absent, i.e., their amplitudes are zero.

Sounds of the human voice and of musical instruments, as well as noises, rustles, etc., are complex, with a large number of harmonics. These harmonics give each sound a certain quality, called *timbre*. A violin and a piano can give a sound of the same strength and frequency, but the sound will be different in each case because it will have different harmonics, and hence a different timbre.

The ultimate link of any broadcast system is the human ear — the actual receiver of the radio programme. However, the ear can not detect all the sounds. *The range of frequencies audible to the ear extends from approximately 20 to 20,000 cps.* As far as the loudness of sounds is concerned, it also has its low and high limits. Too weak a sound is not detected by the ear, while too strong a sound produces a sensation of pain. The ear distinguishes the sounds not only by their frequency but also by their loudness, but the sensation of loudness is not proportional to the actual change of sound intensity; it seems to a man that the strength of a sound has changed but a little, while actually a great change of sound intensity takes place. The ear does not detect small sound intensity changes (smaller than 25%) at all.

The basic frequencies of human speech extend from about 80 to 1,200 cps, their harmonics reaching 8,000 cps. The frequency range of the basic sounds produced by various musical instruments stretches from 30 to 5,000 cps, the harmonic range reaching frequencies as high as 15,000 cps. An ideal reproduction of speech and music calls for the transmission of the 30-15,000 cps sound range, but this is difficult to perform because of technical limitations, and the frequency range of an average broadcast lies within the limits of 50-5,000 cps.

64. MICROPHONE AND EARPHONE

The basic devices of wire telephone communication — the microphone and the earphone — also play an important role in radio communication. A circuit diagram of electrical transmission of sound over wires is given in Fig. 119, which also shows the construction and operation of the carbon microphone and electromagnetic earphone.

A carbon microphone consists of a metal case *C* (Fig. 119*a*) holding a thin carbon diaphragm *CD* and an inner grooved carbon back *CB*. Carbon powder *CP*, held by felt ring *FR*, is placed between the diaphragm and the carbon back. The latter is insulated from the microphone case and is provided with terminal screw *TS*.

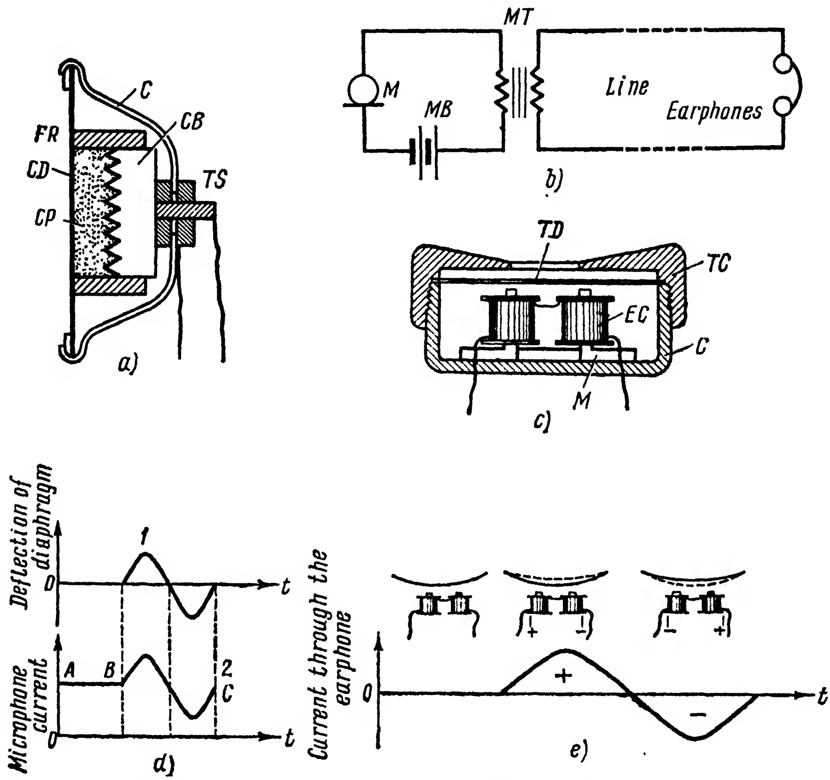


Fig. 119. Construction and operation of carbon microphone and electromagnetic earphone.

The current from the microphone battery *MB* (Fig. 119*b*) flows through the carbon powder, which has a resistance of several tens or hundreds of ohms, and through the series-connected primary winding of microphone transformer *MT*, the step-up ratio of which may be from 1 : 20 to 1 : 100. The secondary winding of this transformer is connected to a line terminating with an earphone.

The earphone case *C*, made of metal or plastic, contains a permanent magnet *M* (Fig. 119*c*). This magnet is provided with pole pieces made of soft steel and carrying electromagnetic coils *EC* comprised of a large number of turns of thin wire, the resistance of which can reach several thousand ohms. A thin tin diaphragm *TD*

is placed over case C, and a small air gap left between it and the pole pieces of the magnet. The diaphragm is pressed to the case edge by threaded cap TC, which has a hole in the centre.

The electrical processes taking place in the microphone and earphone are pictured in Fig. 119 *d* and *e*. When the microphone receives no sound, the resistance of its carbon powder and the current flowing through the powder are both constant (section AB of curve *2* in Fig. 119*d*). Hence under such conditions no voltage is induced in the secondary winding of the microphone transformer.

Sounds reaching the microphone make the diaphragm vibrate and follow them, in accordance with their characteristics (curve *1* in Fig. 119*d*). For the sake of simplicity the drawing presents a single and sinusoidal sound acting upon the microphone diaphragm. The vibrating diaphragm alternately presses and decompresses the carbon powder, thus alternately decreasing and increasing the resistance of the carbon pile. This, in turn, varies the current flowing through the microphone, the current assuming a pulsating character and following the vibrations of the diaphragm (section BC of curve *2* in Fig. 119*d*). This current has d.c. and a.c. components.

The latter component, passing through the primary winding of the microphone transformer, induces a stepped-up voltage in the secondary winding, and the characteristics of this voltage will also correspond to the vibration of the diaphragm, i.e., correspond to the character of sound reaching the microphone. It is this voltage that produces alternating current in the communication line, this current reaching the earphone and making it reproduce the sound which has reached the microphone.

As follows from the above description, a microphone is a device for converting sound into electrical oscillations. Conversely, an earphone converts electrical oscillations into sound. When no current flows through earphone coils, the diaphragm is attracted in the direction of the magnet pole pieces and is slightly bent in (Fig. 119*e*).

The strength of the permanent magnet varies when current passes through its coils. A positive half-wave of alternating current strengthens the magnet, because it sets up a magnetic field which adds up with the field of the magnet; this makes the diaphragm bend in farther. A negative half-wave produces the opposite effect and the diaphragm moves away from the magnet. As a result, the diaphragm vibrates in both directions from its initial position, the frequency of such vibrations being determined by the frequency of the alternating current passing through the earphone coils. Such a diaphragm reproduces the vibration of the microphone diaphragm and sets up a sound wave, which can be heard if the earphone is held close to the ear.

If the earphone did not employ a permanent magnet, the diaphragm vibrations would be weaker and each half-wave of the alter-

nating current would cause an attraction of the diaphragm in the direction of the current-carrying coil. In this case the diaphragm would deflect only in one direction from the position of its equilibrium, and the reproduced frequency would be double the actual frequency transmitted by the microphone. This is remedied by the permanent magnet installed in the coil; the magnet eliminates the frequency-doubling effect and at the same time increases the sensitivity of the earphone.

Nevertheless, a microphone and earphone of the described type set up considerable distortion in the sounds being transmitted, because they do not transmit many harmonics belonging to complex sounds and, besides, produce unrequired vibrations of their own. The diaphragm-type carbon microphone shown above and called "button microphone" gives large distortion, and its transmission is accompanied by rustles and hissing. It is also prone to "baking", when excessive values of current flowing through it make the carbon particles adhere to each other, welding them together with tiny sparks passing between them. Consequently, a "baked" microphone stops changing its resistance when being spoken in. It is true that the "baking" may be eliminated by shaking the microphone, but this type of microphone possesses still another disadvantage—it is sensitive to humidity and can be put out of order when operated in moist atmosphere. High value of e.m.f. developed by the carbon microphone (0.2-0.4 v) and a low value of the voltage of the microphone battery (about 1-4 v) are the advantages of such a microphone, which is generally used for the transmission of speech only, though not for broadcasting.

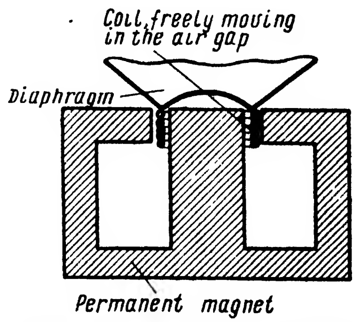


Fig. 120. Construction of a dynamic microphone

The widely used electrodynamic, condenser and piezoelectric types of microphones produce very little distortion.

Electrodynamic or induction microphones are manufactured in two type varieties: *moving-coil* microphone and *ribbon* microphone. The moving-coil microphone is supplied with a thin aluminium diaphragm which carries a small light coil placed in the space between the poles of a strong permanent magnet (Fig. 120). When acted upon by sound waves the diaphragm vibrates, making the coil oscillate and cross the magnetic lines of force. E.m.f. set up in the coil is then fed to an amplifier. The ribbon microphone makes use of a corrugated accordion-shaped aluminium foil stretched in the air gap between the poles of a strong magnet. Sound waves vibrate the ribbon, the latter oscillating in the magnetic field and generating alternating e.m.f.

Electrodynamic microphones develop a very low value of e.m.f. and require high-gain amplifiers. A step-up transformer is usually installed in such a microphone. It differs from the carbon microphone in that it produces no internal noises of its own.

A condenser microphone is, in effect, a peculiar type of air-dielectric capacitor in which one of the plates is made of a thin metal sheet and is located close to the other (solid) capacitor plate. The thin capacitor plate — the diaphragm — is capable of vibrating under the action of sound waves. Such a vibration varies the thickness of the capacitor dielectric (the air gap), and correspondingly makes the capacitance fluctuate. This type of microphone is connected in series with a voltage source E and a high resistance R into the grid circuit of an amplifier (Fig. 121). Fluctuations of its capacitance give rise to charging and discharging currents, which, passing through resistance R, develop a.c. voltage across the resistor. Actually, an increase of the capacitance creates an additional charge of the microphone, making the battery supply a certain amount of charging current, while a decrease of the capacitance produces the opposite effect—the capacitor will partly discharge and the current will flow from it to the battery.

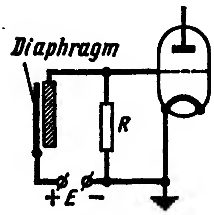


Fig. 121. Circuit connection of a condenser microphone

The alternating voltage built up across resistance R is very small and must be greatly amplified to be of practical use. Since various types of electrical noise reaching the input of a high-gain amplifier are likely to result in the same noise level as the microphone signal level at the input of the amplifier, this is precluded by placing the first amplifier stages right into the case of the condenser microphone.

This microphone has a very low distortion but its quality is nevertheless not as good as that of the electrodynamic microphone.

The condenser microphone, because of the high resistance of its circuit and the high voltage it employs, can develop internal noise as a result of stray interfering currents appearing due to minor imperfections of the microphone insulation.

The *piezoelectric microphone*, also called the crystal microphone, operates by *piezoelectric effect*. The word "piezoelectricity" denotes "electricity obtained by means of pressure". In some types of crystals the piezoelectric effect is observed when opposite edges of the crystal, subjected to mechanical pressure, develop electrical charges of opposite polarity and of equal value. In this case the magnitude of the electrical charges is proportional to the pressure, and the direction in which the pressure is applied determines the polarity.

Piezoelectric devices used in radio generally employ artificial crystals of chemical substance known as *Rochelle salt*, out of which rectangular plates are cut in a certain way.*

The opposite sides of a plate thus cut out are covered with a thin metal layer, forming a *piezo cell* with two terminals. If a piezo cell is subjected to pressure applied in a certain direction, one of its terminals develops a positive potential and the other a negative potential (Fig. 122*a*). The polarity will be reversed if the direction of crystal deformation—i.e., the direction of pressure—is changed (Fig. 122*b*). This phenomenon is known as the *direct piezoelectric effect*.

When a piezo cell is subjected to the action of sound waves, it alternately expands and contracts, thus developing an alternating e.m.f. across its terminals, the magnitude and frequency of such e.m.f. following the characteristics of the applied sound. This is the principle of operation of the piezoelectric microphone.

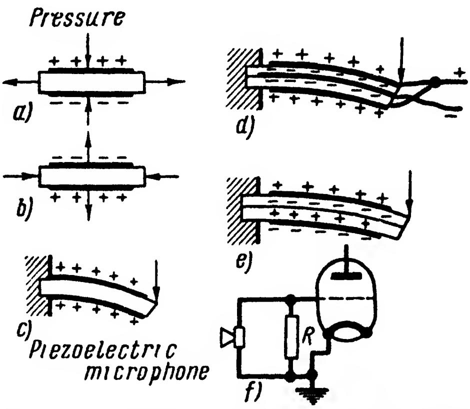


Fig. 122. Piezo cell functions and circuit connection of a piezoelectric microphone

In practical constructions piezo cells are usually fixed either by one or by both ends and, when acted upon by sounds, they do not simply expand and contract but rather bend, as shown in Fig. 122*c*. But in this case one half of the cell (the lower half in Fig. 122*c*) contracts, while the other half expands.

This brings about a situation when the charges appearing on the terminals of the cell have the same polarity. In order to obtain opposite polarities in a piezo cell subjected to bending, special *bimorphous piezo cells* are used. Such cells constitute, in fact, two ordinary piezo cells glued together (Fig. 122*d*). The outer terminals of a bimorphous cell are connected in parallel and develop one polarity, while the common inner terminal of the same cell develops the opposite polarity. If when the cells are glued together the sides of one of them are reversed, a bimorphous cell with series connection of the terminals will be obtained. Such a cell, shown in Fig. 122*e*, gives a doubled voltage output.

Some piezoelectric microphones employ so-called "sound cells", consisting of several bimorphous cells directly acted upon by sound

* Certain radio devices use piezoelectric substances other than the Rochelle salt, e.g., ammonium phosphate or ceramic barium titanite.

waves. Another variety of piezoelectric microphone uses a thin metal diaphragm acted upon by the sound waves and transmitting the vibrations to a piezo cell. The piezoelectric microphones possess considerable capacitance (several hundred picofarads).

It should be kept in mind that the Rochelle salt is very weak mechanically, melts at 63° C and can be used only at temperature not in excess of 40° C. This salt readily dissolves in water; therefore piezo cells must be protected from moisture.

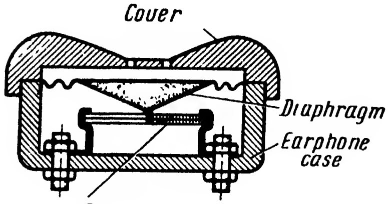


Fig. 123. Construction of a piezoelectric earphone

The crystals of certain minerals, such as quartz, tourmaline and some others, also possess piezoelectric properties. Unlike the Rochelle salt crystals, they are very stable both mechanically and thermally, but their piezo effect is much weaker. They are therefore not employed by electroacoustic devices but find application in the frequency control of valve oscillators and transmitters (see Chapter VIII):

Piezoelectric microphones have very low distortion, give a high output and are connected directly to the grid of amplifier valves (Fig. 122*f*). To prevent electrons from accumulating on the grid, resistor R must be connected between the grid and cathode in such circuits.

The piezoelectric phenomenon is utilised not only by microphones but also by earphones. Piezoelectric earphones make use of the so-called *inverse piezoelectric effect*. This is as follows: if e.m.f. is applied from an external source to the terminals of a piezo cell, the cell will either expand or contract in the direction determined by the polarity of the applied electrical charge. A bimorphous cell will bend under the influence of the electrical charge. The higher the difference of electrical potentials, the more pronounced will be the crystal deformation. If alternating e.m.f. is applied to a piezo cell, the cell will oscillate.

The direct piezoelectric effect constitutes a conversion of mechanical energy into electrical, while the inverse piezoelectric effect represents the opposite process. Therefore piezoelectric electroacoustic devices are called *electromechanical convertors*.

The construction of a piezoelectric earphone is shown in Fig. 123. It employs a bimorphous cell fixed by three of its corners, the fourth being attached to an aluminium diaphragm. An ohmic resistance is usually connected in series with the cell. In comparison with electromagnetic earphones, a piezoelectric earphone has lower distortion, higher sensitivity and much higher impedance. Its disadvantage is the instability of the Rochelle salt noted above. Piezoelectric earphones should be protected from moisture. Avoid dismantling this type of earphone, because they are produced with a sealed cover.

65. LOUDSPEAKERS

Like an earphone, a loudspeaker converts low-frequency alternating-current energy into the energy of sound waves. An earphone gives weak signals because the amplitude of vibrations of its diaphragm is low and the diaphragm is small. If a considerable a.c. voltage is applied to an earphone, the device will distort the sound and rattle. Loudspeakers are used for the reproduction of high-power audio oscillations and differ greatly from earphones in design. A good loudspeaker not only delivers high-power audio output but must also faithfully reproduce sounds of different frequencies. The last quality, called fidelity, is difficult to obtain, and as a rule most loudspeakers reproduce certain frequencies better than others, and do not reproduce some frequencies at all.

In acoustic design loudspeakers are classified into *cone loudspeakers and horn loudspeakers*. In the first type the sound is radiated into the air by a large cone-shaped paper diaphragm, attached to an electromechanical drive unit which sets the cone into vibration. In horn loudspeakers, as in an ordinary gramophone, the sound is radiated from the drive unit by means of a horn. These have directional properties and are mainly used for outdoor sound diffusion and public address installations, as well as in motion picture houses. Ordinary radio receivers employ cone loudspeakers exclusively.

Electromagnetic loudspeakers. These loudspeakers are now obsolete and are no longer made. The most common loudspeaker of this type was called "Record".

Electrodynamic loudspeakers. These devices, often called simply "dynamic speakers", are the basic type of modern loudspeaker.

Fig. 124 shows the construction of a dynamic speaker. The speaker employs magnet *M* with a closed magnetic circuit, shaped like the Russian letter Ш (Fig. 124*b*) or 2 cylinders. A light-weight voice coil *VC* is placed in the air gap around the tip of the central leg of the electromagnet. The voice coil is supplied with audio frequency current. The interaction of the alternating magnetic field set up by these currents in the voice coil, and of the constant field of the magnet makes the voice coil vibrate, moving up and down in the air gap around the tip of the central leg.

The speaker employes a centering washer *CW* to assure exact positioning of the voice coil in the centre of the air gap. This washer is made of flexible material and is provided with shaped cut-outs, which further increase its elasticity (Fig. 124*c*). The washer edges are glued to the edges of the voice coil frame. The washer is fixed by its centre to the central leg of the magnet in such a way that they have a common axis. The voice coil is glued to paper cone *C* (Fig. 124*d*), whose edges are provided with an accordion-type attachment secured to the ring-shaped frame of the loudspeaker.

The voice coil of the dynamic speaker has a low resistance; hence this type of loudspeaker is a low-impedance device. Therefore, dynamic speakers are always connected to radio receivers, amplifiers and sound diffusion systems through step-down transformers, whose primary windings contain a large number of turns. On audio frequencies the impedance of such a transformer is equal to several thousand ohms (this point is discussed in greater detail in Chapter VII). The transformer is frequently installed right on the loudspeaker frame. Formerly dynamic speakers with electromagnets instead of magnets were made, but these are no longer manufactured.

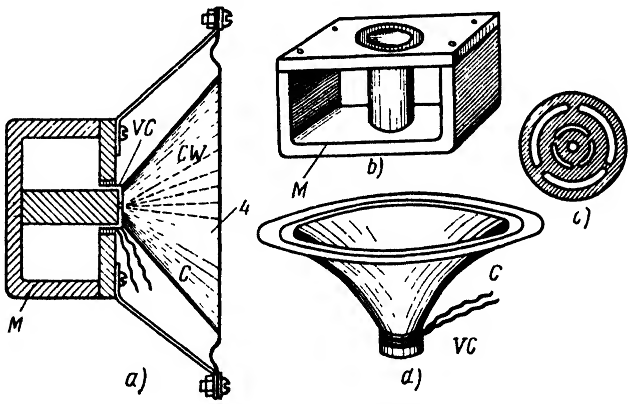


Fig. 124. Construction and components of an electrodynamic loudspeaker: *a)* cross-sectional view of the loudspeaker, *b)* magnetic system, *c)* centering washer, *d)* speaker cone with voice coil

It became possible to make dynamic speakers using permanent magnets instead of field electromagnets only after the discovery of special magnetic alloys, out of which small but powerful permanent magnets are made.

Good reproduction of sounds (and particularly of low-frequency sounds) requires that the loudspeaker is provided with an acoustic screen, which is actually a wooden board or box in which a round hole is cut out for the loudspeaker. In a radio receiver the receiver cabinet plays the role of acoustic screen.

Permanent-magnet dynamic speakers rated at 0.2-0.5 va* are specifically designed for operation from sound-diffusion lines. Such speakers are enclosed in plastic cabinets together with transformers and gain controls. Many other varieties of dynamic speakers, ranging in power up to 10 va, are now in production.

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* Power consumed by loudspeaker is not pure active power and is, therefore, measured not in watts but in volt-amperes.

Horn loudspeakers employ a thin metal cone-shaped diaphragm instead of paper cone. Such a diaphragm has much smaller dimensions than the cone of an average radio receiver. Permanent-magnet horn loudspeakers are manufactured with various power ratings, extending from 10 va to 100 va.

Most cone loudspeakers are capable of reproducing sound frequencies from 100 to 6,000 cps. In horn loudspeakers, this frequency range is somewhat narrower and extends from about 200 to 4,000 cps. Both types of loudspeaker give satisfactory performance within the specified frequency limits.

A single loudspeaker cannot reproduce faithfully all the frequencies of the sound range. A small cone or horn speaker is capable of good reproduction of the higher audio frequencies, but its performance falls off in lower frequencies of the sound range. On the other hand, a large speaker reproduces the low frequencies well, its performance falling off on the higher frequencies. Where high-fidelity reproduction is required, two loudspeakers are employed. One of them is specifically designed to reproduce only the low and middle frequencies of the sound range, the other—only the middle and high frequencies. Such a speaker combination, used by motion picture sound equipment, reproduces a frequency range extending from 40 to 9,000 cps. A combination like this always uses a filter network, an example of which is shown in Fig. 125. Such a network actually consists of two filters, one of which (L_1C_1) does not pass the higher audio frequencies, e.g., above 1,000 cps, but passes only the frequency band of 40-1,000 cps, directing these frequencies to the large (low-frequency) loudspeaker. On the other hand, filter L_2C_2 does not pass the lower frequencies and feeds the high-frequency speaker only with frequencies higher than 1,000 cps.

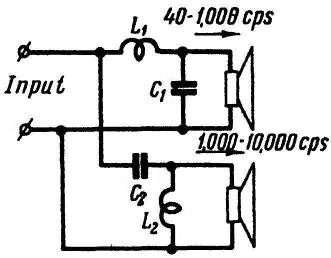


Fig. 125. A combination of loudspeakers and associated filter network

Piezoelectric loudspeakers. The principle of design and operation of the piezoelectric loudspeaker is the same as that of the piezoelectric earphone. Loudspeakers of the given type are provided with bimorphous piezo cells which drive the speaker cone. Such loudspeakers reproduce the frequency range between 250 and 3,500 cps. The modern design version is provided with a step-up transformer or autotransformer and a gain control, represented by a variable resistor connected in series with the loudspeaker.

66. GRAMOPHONE PICKUPS

A pickup is a device for the electrical reproduction of sounds recorded on gramophone records. As a record rotates, the pickup needle moves along its grooves and, following the groove modulations, oscillates in accordance with the recorded sounds, both in frequency and in shape. The mechanical oscillations of the needle are converted in the pickup into electrical oscillations, which are applied to the input of an amplifier.

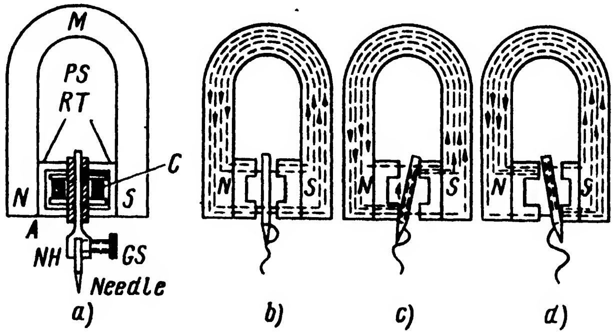


Fig. 126. Construction and principle of operation of an electromagnetic pickup

The construction of an electromagnetic pickup is shown in Fig. 126*a*. The pickup is comprised of permanent magnet *M* provided with U-shaped pole shoes *PS*. Miniature armature *A*, carrying needle-holder *NH* with grub screw *GS*, is positioned in the air gap between the two pole shoes and is capable of vibrating when a vibratory force is applied to the needle inserted in the needle-holder. The armature, together with rubber tube *RT* slipped over it, is placed in the centre hole of coil *C*. This coil consists of a large number of turns; its cross-section is shown in Fig. 126*a*.

When the needle follows a smooth (unmodulated) groove, the armature remains in its neutral position, and there is no magnetic flux along the armature, i.e., through the coil axis (Fig. 126*b*). Encountering a modulated section of the groove, the needle deflects the armature from the neutral position (Fig. 126 *c* and *d*).

This changes the clearances between the armature and the pole shoes, some parts of the air gap decreasing and other parts increasing, as a result of which part of the magnetic flux passes along the armature in one direction or the other. The varying magnetic flux induces alternating e.m.f. in the coil with an amplitude reaching several fractions of one volt. Electromagnetic pickups are capable of reproducing a frequency range extending approximately from 50 to 5,000 cps.

Besides this type of pickup, piezoelectric pickups are also widely used. The construction of this pickup is simpler than that of an electromagnetic pickup. The design principle of the electromechanical part of a piezoelectric pickup is shown in Fig. 127*a*. Here a bimorphous piezo cell, fixed at one end, carries a needle-holder at the other. The piezoelectric pickup reproduces a wide frequency range — from the lowest frequencies up to about 10,000 cps. It develops an e.m.f. as high as 1-1.5 volts.

This type of pickup is usually connected through a gain control directly to the grid of an amplifier valve (Fig. 127*b*) without an intermediate transformer. It should be remembered that the piezo cell does not conduct electrical current; therefore when such pickup is used without a gain control, a 0.5-megohm resistor must be connected between the cathode and control grid of the input valve (Fig. 127*c*).

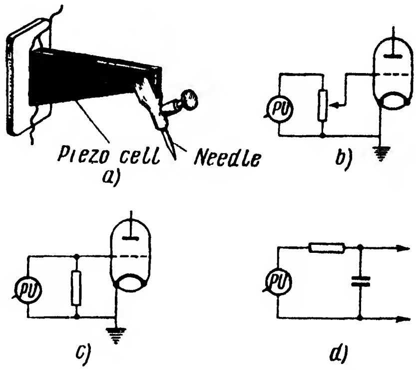


Fig. 127. Principle of construction of a piezoelectric pickup, and its connection to the circuit

Gramophone records develop a certain level of internal "hissing", particularly pronounced when a piezoelectric pickup is employed. A hiss-suppressing filter, lowering the gain on higher audio frequencies, is generally incorporated in the input circuit just described (Fig. 127*c*).

67. THE DECIBEL

Sound intensity (sound volume) I_0 is the energy transferred by a sound wave in one second through a 1 cm² area perpendicular to the direction of propagation of the sound wave. In other words, the intensity of sound is the power contained in 1 cm² of sound-wave cross-section. This is why the sound intensity, otherwise known as sound volume, is sometimes also referred to as *sound power*. Units of measurement of sound intensity are the watt and microwatt per square centimetre (w/cm² and μw/cm²). Every sound is also characterised by the value of *sound pressure* p, i.e., by that additional pressure which, apart from atmospheric pressure, is created by the sound wave in a given part of space. The value of sound pressure is variable and is therefore expressed as a peak value p_m or as an effective value p, like the alternating electric current values. Sound pressure is measured in bars, one bar being approximately equal to the force of pressure of 0.001 g per cm². There is a quadratic relationship between the sound intensity and sound pressure: when the sound pressure increases by 2, 3, 4, etc., times, the sound intensity increases by 4, 9, 16, etc., times.

Sound loudness is the value characterising the response of the human ear to a sound. For the purpose of comparing sounds of different loudness, logarithmic units of measurement have been introduced. Such units give an accurate measurement of the response of the ear to a sound. A measurement unit of this type, used in radio engineering, is known as the *decibel (db)*.

The human ear is most sensitive to the sounds of the middle part of the audible frequency range. In electroacoustics 1,000 cps is taken as the reference frequency of this part of the range. The weakest sound audible to the ear on this frequency has an intensity (I_0) equal to 10^{-16} w/cm² ($10^{-10}\mu$w/cm²). This corresponds to a sound pressure of 0.0002 bars. The values of $I_0 = 10^{-16}$ w/cm² and $p_0 = 0.0002$ bars are taken as reference values corresponding to the *zero level* of sound. In measurement practice sounds of all frequencies are referred to this zero level, which is expressed as zero decibels (0 db) in the system of logarithmic units of measurement.

If the sound intensity is changed by 10, 100, 1,000, etc., times, such changes correspond, respectively, to a 10, 20, 30, etc., decibels change in the logarithmic system of measurement units. A 1-decibel change corresponds to changing the intensity of sound by about 1.25 times, i.e., by 25%.

Incidentally, this is the smallest change of sound level perceptible to the ear. On 1,000 cps, the zero level of sound intensity coincides with the zero level of loudness, such a level being known as the threshold of audibility. Sounds of frequencies other than 1,000 are associated with other zero levels, because the ear sensitivity is different on various frequencies. For instance, a 40-cps sound has the same sound intensity zero level as a 1,000-cps sound, but the zero level of its loudness corresponds to intensity of 10^{-10} w/cm², i.e., corresponds to a sound whose power is one million times as large and has a level of 60 db.

This considerably complicates all calculations and measurements connected with the sounds of different frequencies. To simplify matters, it has been agreed that the level of loudness of a given sound of any frequency is considered the same as the intensity level of a 1,000-cps sound if the latter sound has the same loudness as the unknown sound.

Table 2 gives the complete range of loudness levels, this range extending from 0 db to 130 db.

Studying the above table and considering the fact that the human ear can distinguish a change of sound intensity only when its level is at least one decibel, it becomes evident that the whole loudness range is covered by 130 sounds of different loudness. In practice we seldom deal with weak sounds close to the threshold of hearing or with strong sounds that approach the threshold of pain.

A whisper or the softest-playing violin gives a level of about 30 db, while a fortissimo of a large orchestra corresponds to a 90-db level of loudness. Hence the loudness range of sounds usually perceived by the ear is approximately 60 db. The range of loudness levels or the difference between the levels of the strongest and weakest sounds is called the *dynamic range*.

Table 2 offers a comparison of sounds of different loudness, making it possible to transpose sound pressure and sound intensity into decibels and vice versa. The columns listing the decibels and the relationships of sound pressures or sound intensities help to compare the levels of any sounds (not necessarily in reference to the zero level). Thus an average conversation at a distance of 1 metre creates a sound pressure of 0.2 bars and has a loudness level of 60 db in reference to the zero level. However, when compared to a weak sound (whisper) having a level of 30 db, the level of such conversation becomes equal to 30 db, because $60 - 30 = 30$. As seen from the table, the value of 30 db corresponds to 31.6 in the column of sound pressure relationships. It also corresponds to 1,000 in the column of sound intensity relationships. Therefore the level of an average conversation is higher than the level of a whisper by 31.6 times in respect to the sound pressure, and by 1,000 times from the point of sound intensity.

Radio engineers frequently resort to decibels not only when it is required to express an increase or decrease of loudness or sound intensity, but also when it is desirable to calculate amplification or attenuation of voltage or power levels in alternating current circuits.

Table 2 also offers assistance in such calculations; the column of sound pressure relationships gives the relationship of voltages, and the column of sound intensity relationships that of power levels.

Table 2

Loudness level in db	Sound pressure in bars	Sound intensity in $\mu w/cm^2$	Relationship of sound pressures (or voltages)	Relationship of sound intensities (or powers)	Brief characteristic of sound
0	0.0002	10^{-10}	1	1	Threshold of hearing
10	0.00065	10^{-9}	3.16	10	Soft whisper at a distance of 1 m
20	0.002	10^{-8}	10	10^2	Quiet garden
30	0.0065	10^{-7}	31.6	10^3	Whisper at a distance of 1 m or softest violin playing
40	0.02	10^{-6}	100	10^4	Soft music or noises in an average residence
50	0.065	10^{-5}	316	10^5	Low-level operation of a loudspeaker
60	0.2	10^{-4}	10^3	10^6	An average conversation at a distance of 1 m
70	0.65	10^{-3}	3.16×10^3	10^7	Noise inside a tram-car
80	2	10^{-2}	10^4	10^8	High-level operation of a loudspeaker or a noisy street
90	6.5	10^{-1}	3.16×10^4	10^9	Automobile horn or fortissimo of a large orchestra
100	20	1	10^5	10^{10}	Riveter or an automobile siren
110	65	10	3.16×10^5	10^{11}	Pneumatic hammer
120	200	10^2	10^6	10^{12}	Airplane engine at a distance of 5 m or a strong thunderclap
130	650	10^3	3.16×10^6	10^{13}	Sensation of pain; sound is no longer perceptible

The sensitivity of a microphone is expressed in terms of e.m.f. developed by the microphone acted upon by 1-bar sound pressure. Thus the measurement unit of microphone sensitivity is one millivolt per bar (mv/bar).

Electrodynamic microphones, for instance, have sensitivity of about 0.1 mv/bar, while the sensitivities of condenser and piezoelectric microphones vary from 0.5 to 5 mv/bar. Sometimes the sensitivity of a microphone is expressed in decibels in relation to some reference sensitivity, the latter being considered as the zero level.

The dependence of microphone sensitivity upon frequency is called the *frequency-response characteristic* of a microphone (Fig. 128*a*). Unequal response of a microphone to different frequencies, i.e., non-linearity of the frequency-response characteris-

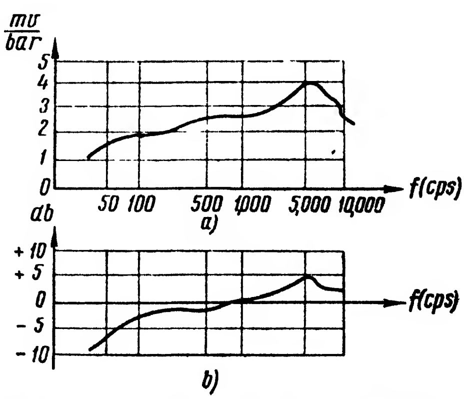


Fig. 128. Various means of presenting frequency-response characteristics of a microphone

tic, is an indication that the microphone introduces so-called *frequency distortion* in its circuit. Such distortion is usually expressed in decibels. For instance, if the frequency-response curve of a certain microphone has a 10-db irregularity, a reference to Table 2 will show that the voltage developed by the microphone varies by as much as 3.16 times for sounds of different frequencies but of the same pressure. Sometimes the frequency-response characteristic of a microphone is given as a curve showing the sensitivity variation in decibels, as related to some average sensitivity or to the sensitivity on 1,000 cps. If the sensitivity of the microphone under test happens to be above average, the sensitivity deviation is considered as positive; the deviation is considered negative when the microphone sensitivity is below the average (Fig. 128*b*).

The sensitivity of loudspeakers is expressed in terms of sound pressure in bars, this pressure being created at a definite distance from the loudspeaker when the latter is connected to a 1-volt source of alternating voltage (this gives bars per volt). Sometimes the loudspeaker sensitivity is not referred to the 1-volt source but to 1-va of power fed to the loudspeaker under test. The dependence of loudspeaker sensitivity upon frequency is the frequency response characteristic of a loudspeaker.

Irregularities occurring in such characteristic, i.e., unequal response on various frequencies, is usually expressed in decibels and characterises frequency distortion in the reproduction of sound by the given loudspeaker. For instance, room-type speakers, designed for the reproduction of sound diffusion programmes, create frequency distortion not in excess of 20 db in the 150-6,000 cps frequency band. This means that the sound pressure in this frequency band does not change by more than 10 times when the same value of voltage of different frequencies is applied to the terminals of the loudspeaker.

68. QUESTIONS AND PROBLEMS

1. What are the 5th and the 7th harmonic frequencies of a complex sound if its basic frequency is 250 cps?
2. If the amplitude of a sound has been doubled, will the sound seem twice as loud to the ear?
3. The eighth harmonic of a complex sound has a frequency of 3,000 cps. What is the basic frequency of such sound?
4. Is it correct to say that a carbon microphone converts sound oscillations into audio-frequency alternating currents?
5. What is the function of a microphone transformer?
6. An earphone can operate as a microphone. Explain this.
7. If the magnets of an earphone and of a pickup are demagnetised, will these two devices function?
8. What is the function of the permanent magnet in an earphone?
9. What are the advantages and disadvantages of a carbon microphone as compared to other types?
10. What types of microphone do not require direct-current supplies?
11. What kind of current flows through the circuit of a condenser microphone—alternating or pulsating current?
12. What is the piezoelectric effect?
13. How would you connect a gain control to a loudspeaker?
14. What is dynamic range?
15. What is the frequency distortion taking place in microphones and loudspeakers?
16. What is the frequency-response characteristic of a loudspeaker?
17. Under the condition of equal sound pressure, a certain microphone gives a voltage output from 0.05 to 0.5 mv on different frequencies. Express the frequency distortion in decibels.
18. By how many decibels will the loudness of a sound decrease if the voltage applied to the loudspeaker is changed from 30 to 3 volts?
19. Why is it that Rochelle salt, not quartz, is used in piezoelectric earphones and microphones?

CHAPTER VII

LOW-FREQUENCY AMPLIFIERS

69. THE BASIC PARAMETERS OF AMPLIFIERS

Every amplifier has *input terminals*, to which is fed *the voltage to be amplified*, *output terminals*, *across which the amplified voltage appears*, and power facilities, supplying the filament, anode and grid circuits of the amplifier.

The input circuit of an amplifier is connected to the generating device whose low-frequency voltage the amplifier is to amplify. Such generating device may be a microphone, gramophone pickup, receiver output, photocell (in motion picture sound equipment), etc.

The output circuit of the amplifier can feed a loudspeaker, or a line terminating with loudspeakers and earphones, or any other device consuming the power of the amplified electrical oscillations.

Let us now get acquainted with the basic values characterising the operation of an amplifier.

Amplification factor of an amplifier. The ratio of voltage U_o, developed by the output circuit of an amplifier, to voltage U_i, applied to the input of the amplifier, is called the amplification factor k of the amplifier and is represented as follows:

$$k = \frac{U_o}{U_i}.$$

The amplification factor (or the gain) of an amplifier depends upon the amplification factors of the individual stages making up the whole of the amplifier circuit. If these amplification factors are denoted, respectively, by k_1, k_2, k_3, etc., the following relation may be written:

$$k = k_1 k_2 k_3.$$

For example, if an amplifier has three stages with the following amplification factors: $k_1 = 20$, $k_2 = 20$ and $k_3 = 5$, the overall gain of the whole amplifier will be given by:

$$k = 20 \times 20 \times 5 = 2{,}000.$$

This means, that if a 10-millivolt signal U_i is applied to the input of such amplifier, the resultant output voltage will be given by: $U_0 = 2{,}000 \times 10 \times 10^{-3} = 20$ volts.

Output power P_0. The output power P_0 of an amplifier is the power of low-frequency alternating current developed by the amplifier in its load resistance. In low-power amplifiers, P_0 may be equal to a fraction of one watt, in medium-power amplifiers—to several watts or several dozens of watts, while a high-power amplifier is capable of developing an output power expressed in hundreds of watts and even in kilowatts. The output power level is always stipulated for normal operating conditions of every amplifier, i.e., when normal required voltage is applied to the input circuit of the amplifier.

Every amplifier is capable of giving a power output in excess of the normal rated value. However, such overloaded operating condition of an amplifier results in higher distortion. And, besides this, excessive values of alternating voltage in the output stage of an amplifier can cause failures of amplifier components (breakdown of insulation in capacitors or transformers, failure of valves, etc.).

Frequency range. Depending upon the purpose served by an amplifier, the range of frequencies evenly amplified by it may be either broad or narrow. When the amplifier is designed for speech amplification only, its frequency range lies between about 200 to 2,000 cps. High-fidelity amplifiers, used for the amplification of musical programmes, have a frequency range extending from 50 to 10,000 cps.

Amplifier distortion. When electrical oscillations are amplified, the following types of distortion arise in the amplifier circuits.

1. *Frequency distortion.* The broader the frequency range of oscillations normally handled by an amplifier, the smaller is the distortion of such oscillations during their amplification by the circuits of the amplifier. An ideal amplifier is supposed to amplify equally well the oscillations of any frequency included in the frequency range for which the amplifier is designed. However, such ideal amplifier is but an assumption and all practical amplifiers amplify unequally the oscillations of different frequencies.

This disturbs the fidelity of reproduction and gives rise to what is known as *frequency distortion*, also called linear distortion. A *frequency characteristic*, representing the dependence of amplification factor k upon frequency f of the oscillations being amplified, serves as an index of the magnitude and character of frequency distortion in an amplifier. When such frequency characteristics are plotted, it is customary to mark off the frequency on logarithmic scale along the X-axis (Fig. 129). (The logarithmic scale means that the divisions become smaller as the frequency increases; such a scale is resorted to because it would be difficult to "squeeze in" the whole wide range of audio frequencies into the ordinary linear scale, where all the divisions are of the same size). Along the Y-axis is plotted

not the numerical value of the amplification factor k, but the ratio of this factor at a given frequency to the factor at some standard average frequency, which is usually taken to be equal to 400 cps (sometimes, 1,000 cps).

Fig. 129a gives the frequency characteristic of the ideal amplifier designed for the frequency range of 50-10,000 cps. Apparently, the characteristic indicates an unchanging amplification factor for all the frequencies of the range, i.e., an absence of frequency distortion. However, such an ideal amplifier, as has been already noted, is difficult to design; and neither is such a construction necessary inasmuch as the human ear cannot perceive minor changes of sound intensity, within a 25% limit. Therefore, slight deviations of the amplification factor from the average value is quite tolerable on most frequencies.

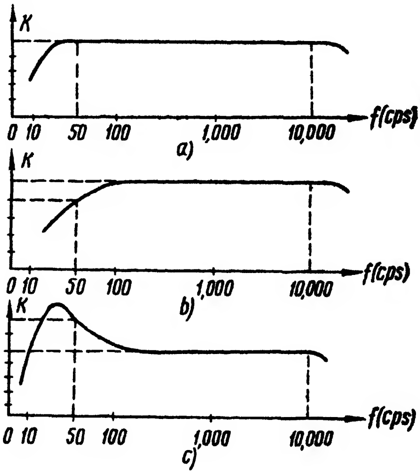


Fig. 129. Various shapes of frequency characteristics

Fig. 129b shows, for instance, the frequency characteristic of an amplifier in which a frequency of 50 cps is amplified 20% less than the middle frequencies. This is called "the attenuation" of lower frequencies.

Fig. 129c illustrates a frequency characteristic which has a lift in its low-frequency region; the amplification factor at 50 cps is 40% higher than at the average frequency. Attenuation, or lifting, of the characteristic can also take place on high frequencies. If such irregularities of a characteristic exceed 20-30%, the distortion becomes noticeable; the ear notices the low level of some of the sounds at the output of the amplifier, or else perceives that certain sounds are reproduced abnormally loud in comparison with the sounds of the average frequencies. Further on, while studying various types of amplifiers, we shall explain the character of frequency distortions arising in each one of them.

The reason for frequency distortion in amplifiers is as follows. Every amplifier circuit employs capacitors and inductance coils. The reactance of these components depends upon frequency. Hence, the amplifier operation varies on different frequencies.

Frequency distortion also arises—and to a considerably greater extent than in amplifiers — in such devices as loudspeakers, earphones, microphones and pickups. Because of this, it sometimes becomes necessary to employ amplifiers with a particular character of

frequency distortion which would compensate (or correct) another type of frequency distortion appearing, for instance, in a loudspeaker driven by such amplifier. If the loudspeaker attenuates the higher frequencies, it becomes desirable to drive it from an amplifier in which the frequency characteristic has a lift at the high-frequency end. This will improve the overall characteristic of the sound system and the reproduction of sound will be accompanied by smaller distortion.

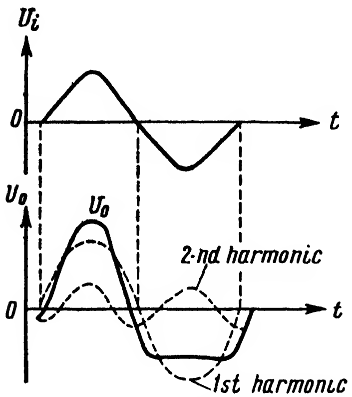


Fig. 130. Non-linear distortion

Some amplifiers employ so-called *boosting controls* and *tone controls* which can be adjusted to vary the frequency characteristic of an amplifier in such a way that frequency distortion occurring in other parts of the sound system is compensated for to the desired degree.

2. *Non-linear distortion.* When a sinusoidal voltage is applied to the input of an amplifier, the amplified signal at the output, as a rule, will not be sinusoidal but will have a more complex form.

Every complex oscillation consists of a series of simple sinusoidal oscillations—the basic oscillation and higher harmonics. Thus, an amplifier, distorting the shape of oscillations under amplification, adds extra harmonics which were not present in the input signal.

Fig. 130 shows a sinusoidal voltage U_i applied to the input of an amplifier and the resultant, distorted, non-sinusoidal output voltage U_0. In this case, the amplifier has introduced the second harmonic into its output circuit. One broken curve in the drawing relates to the output voltage U_0 and shows the useful first harmonic (the basic oscillation) which has the same frequency as the input voltage. The other broken curve shows the undesirable second harmonic, the frequency of which is twice as high as the basic oscillation frequency. The output voltage is the sum of these two harmonics.

Shape distortion of signals, i.e., the addition of extra harmonics to the basic oscillation, is known as non-linear distortion. The presence of such distortion is evidenced by hoarse and rattling sound at the output of an amplifier.

*The coefficient of non-linear distortion k_n is used to evaluate this type of distortion and shows the percentage of all extra harmonics related to the basic oscillation and developed by the amplifier circuit.**

* The coefficient of non-linear distortion is sometimes called the harmonic content.

If the value of k_n is less than 5%, that is, if the sum of all the harmonics developed by an amplifier does not exeed 5% of the first harmonic, or the original signal, the distortion is not noticeable to the ear. When k_n exceeds 10%, the hoarseness of sound and rattling spoil high-fidelity transmissions. Values of k_n greater than 20% are not tolerable at all because they make even the speech quite indistinct.

Non-linear distortion relates not only to the amplification of sinusoidal signals but also to the amplification of complex oscillations encountered in the transmission of speech and music. In the latter case also, the presence of such distortion changes the correct shape of oscillations under amplification, that is to say, harmonics.

Complex oscillations normally consist of harmonics which should be reproduced by an amplifier as faithfully as possible. These harmonics should not be confused with the extra and undesirable harmonics developed in the amplifier as a result of non-linear distortion. Harmonics contained in the input signal are useful harmonics because they account for the correct timbre of sounds.

On the other hand, harmonics generated by the amplifier circuit itself are very undesirable, because they set up non-linear distortion.

The usual causes of non-linear distortion in amplifiers are the non-linearity of valve characteristics, and the appearance of control grid current and magnetic saturation of the cores of transformers and low-frequency chokes employed by amplifiers. All these causes will be discussed later. Considerable non-linear distortion can also be developed in loudspeakers, earphones, microphones and pickups.

3. *Other types of distortion*. *Phase distortion* arises in amplifier circuits because of various reactances employed in such circuits. This type of distortion consists in a change in the phase shifts between oscillations of various frequencies at the output of an amplifier, differing from the original phase shifts at the input. As far as the reproduction of sound is concerned, such phase shifts are of no importance at all because the ear does not respond to them. However, in some other cases—for instance, in the transmission of television programmes—the phase shifts produce undesirable effects.

Apart from the various types of distortion discussed above, every amplifier also introduces a *distortion of its dynamic range*. As a rule, a compression of the dynamic range is observed, which means that the ratio of the strongest to the weakest signals is considerably smaller at the output of an amplifier than a similar ratio at the input. This upsets the naturalness of reproduced sounds. *Expander-circuits* are sometimes provided in amplifiers to eliminate this undesirable effect. Such circuits serve to broaden (to expand), the dynamic range of amplifiers. It should be noted, however, that not only amplifiers but also other electro-acoustic devices have a tendency to compress the dynamic range.

70. VOLTAGE AMPLIFIERS AND POWER AMPLIFIERS

In every multi-stage amplifier, the first few stages, counting from the input stage, constitute the so-called *voltage amplifier*, otherwise known as a *preamplifier*. Preamplifier stages employ low-power valves.

The main object of the preamplifier is the distortionless amplification of the alternating voltage fed to the input of the amplifier. The amplified input voltage then drives the remaining part of the amplifier, called the *power amplifier* or the *output amplifier*. The power amplifier's function is that of amplifying the power of oscillations with minimum possible distortion and of feeding this power to loudspeakers.

It should not be thought, however, that a preamplifier amplifies voltage only. Like any other type of amplifier, it also increases the power of oscillations, but the operating condition of preamplifier stages is so adjusted that the preamplifier gives the maximum magnification of voltage. Doing this, it also amplifies the alternating current fed to it, and, therefore, amplifies the power of oscillations. On the other hand, a stage actually called a power amplifier frequently operates under such conditions that it does not amplify the voltage but magnifies only the power of oscillations.

The stage preceding the power amplifier stage serves the only purpose of "swinging" the power-amplifier-stage grids and is called the *driver*.

All parts of an amplifier contribute some measure of frequency and non-linear distortion. The final, or the output, stage is a particularly active source of distortion, non-linear distortion being most pronounced in this stage. Constructionally, this stage — the power output stage—can be mounted together with the driver stage, or as a separate unit. The power-supply rectifier, too, can be mounted together with such a combination of output-driver stages, or, as in some designs—separately.

71. A TRIODE AMPLIFIER

The principle of operation of an amplifier stage has already been studied (Secs 32 and 35). A more detailed exposition of this point is given below.

Fig. 131*a* pictures a triode amplifier stage. Here, alternating voltage U_{mg} is fed to the grid of the valve from some generator G, the latter being represented by a pickup, a microphone, or a preceding stage of amplification. A load resistor R_a is put in the anode circuit of the triode amplifier stage. An a.c. component of anode current I_{ma}, appearing across R_a, develops the amplified alternating voltage U_{mR}. A high-capacitance capacitor C (several

microfarads) is connected across the anode battery. To the a.c. component of the anode current this capacitor presents but a slight reactance and easily conducts this component, thus obviating the unnecessary loss of alternating voltage in the internal resistance of the anode battery. It may be considered (as far as the a.c. component is concerned) that capacitor C short-circuits, or shunts-out, the battery and connects the bottom end of R_a directly to the cathode of the valve. Hence, the amplified voltage U_{mR} is, at the same time, the alternating voltage U_{ma} appearing across the anode and cathode of the valve. For the sake of simplification, Fig 131 does not show the heater-circuit of the amplifier stage.

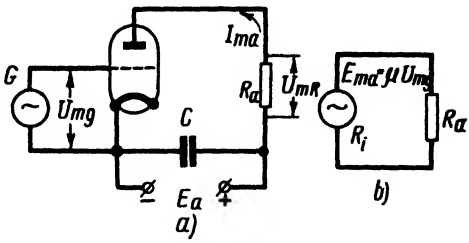


Fig. 131. An amplifier stage and its equivalent circuit

In the circuit diagram of Fig. 131 the generator of input voltage G is shown connected directly to the control grid of the valve. If this generator develops too low a voltage the usual remedy is to connect it to the grid via a step-up transformer (see Fig. 69b). Apart from the alternating voltage the generator, in some cases, also builds up a direct voltage which must be kept away from the grid. The solution in such cases is also the application of an input transformer or of the circuit shown in Fig. 132. In the latter circuit the alternating voltage is applied to the grid through the isolating capacitor C_g which blocks the direct voltage. To minimise the alternating-voltage drop across C_g its capacitance is made several times greater than the grid-cathode input capacitance of the valve. The grid resistor R_g, sometimes called the leakage resistor, is connected between the grid and cathode for leaking off the negative potential set up by the electrons colliding with the grid. If the R_g were not used in the circuit this negative potential would become sufficiently high to cut off the anode current in the valve. The resistance of the R_g is usually several hundred kilohms or higher.

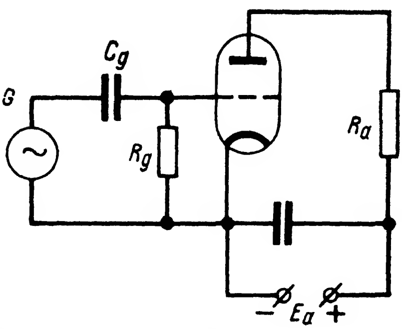


Fig. 132. Capacitor-connection of the input-signal generator

As has already been shown in Chapter IV (Sec. 35), a valve employed in an amplifier stage functions as an a.c. generator possessing internal resistance R_i and giving an anode e.m.f. $E_{ma} = \mu U_{mg}$.

This can be proven as follows. As we already know, the mutual conductance of a valve is given by the steepness of the grid characteristic $S = \frac{\Delta I_a}{\Delta U_g}$. If the grid is supplied with an alternating voltage, then $\Delta I_a = I_{ma}$, while $\Delta U_g = U_{mg}$. Hence, $S = \frac{I_{ma}}{U_{mg}}$ and $I_{ma} = SU_{mg}$. Substituting $\frac{\mu}{R_i}$ for S, we obtain:

$$I_{ma} = \frac{\mu U_{mg}}{R_i}.$$

Treating this formula as the Ohm's law equation we conclude that the value of μU_{mg} stands for the e.m.f. of a generator operating under a condition of short circuit, i.e., under such a condition that the generator output terminals are short-circuited and the entire power given by it is consumed by its internal resistance R_i. This formula holds true for the static operating condition, when no load resistor R_a is employed. However, in an actual amplifier stage, in which the valve operates under the dynamic condition, i.e., when R_a is present, the value of R_a must be added to R_i because both of these resistances are connected in series. In accordance with this, the formula may be finally rewritten as follows:

$$I_{ma} = \frac{\mu U_{mg}}{R_i + R_a}.$$

The formula thus obtained plays an important role in the theory of electron valve circuits and may be considered as the Ohm's law for the a.c. component of anode current of a valve. It should be noted, however, that this formula gives accurate results only when the valve operates on the straight-line part of its characteristic.

On the basis of the derived formula, an *equivalent circuit* of any amplifier stage (from the point of view of alternating current) may be drawn, as pictured in Fig. 131*b*. Here, the valve is shown as an alternator loaded with external resistance R_a. Such an equivalent circuit diagram holds true only for alternating currents and voltages in the anode circuit of a valve. The equivalent circuit diagram of an amplifier stage is very simple. The currents and voltages in such a circuit can be readily calculated from the Ohm's law. It also follows from the circuit, that there is no difference between the voltage across resistor R_a and the voltage across the valve, since this resistor is connected across the generator terminals, i.e., between the anode and cathode of the valve.

Amplification factor (or gain) of any amplifier stage is expressed as k and stands for the ratio of voltages U_{mR} and U_{mg}, as follows:

$$k = \frac{U_{mR}}{U_{mg}}.$$

Using the Ohm's law for the part of a circuit, we have:

$$U_{mR} = I_{ma}R_a.$$

Now, in the above formula, we substitute the following expression for I_{ma} (in accordance with the Ohm's law):

$$U_{mR} = \frac{\mu U_{mg}}{R_i + R_a} R_a.$$

Substituting this expression in the amplification factor formula of the stage, we have:

$$k = \frac{\mu U_{mg} R_a}{(R_i + R_a)U_{mg}}.$$

Dividing the numerator and the denominator by U_{mg}, we obtain the following final result:

$$k = \frac{\mu R_a}{R_i + R_a}.$$

This is the basic formula in the study of operation of various amplifier stages. As seen from the formula, the higher the values of μ and R_a, the higher is the value of k. It is true that factor R_a appears both in the numerator and the denominator, but in the numerator it appears as the multiplier while in the denominator it is but an addendum. Hence, with an increase of the value of R_a, the numerator increases more than the denominator. Such double dependence associated with the value of R_a is explained simply as follows. On the one hand, the greater the value of R_a, the greater is the useful alternating voltage drop across it (R_a in the numerator). However, with an increase of R_a there increases the total resistance of the anode circuit and, therefore, decreases the alternating anode current, which, in turn, causes a reduction of voltage drop across R_a (R_a in the denominator). As far as R_i is concerned, its increase leads to a decrease of k, which is attributed to the greater voltage drop inside the valve.

Example. An amplifier stage employs a valve with the following parameters: $\mu = 20$ and $R_i = 10{,}000$ ohms. Find the value of k for the following values of R_a: 0; 500; 2,500; 10,000; 30,000; 90,000 ohms.

If $R_a = 0$, it is apparent that k is also equal to zero.

When $R_a = 500$ ohms, the result becomes:

$$k = \frac{20 \times 500}{10{,}000 + 500} = \frac{10{,}000}{10{,}500} = 0.95,$$

i.e., there is no amplification as yet because the value of R_a is too small.

When $R_a = 2{,}500$ ohms, the result becomes:

$$k = \frac{20 \times 2{,}500}{10{,}000 + 2{,}500} = \frac{50{,}000}{12{,}500} = 4.$$

There is now some amplification in the stage, but, in comparison with the amplification factor of the valve itself ($\mu = 20$), it is too small.

When $R_a = 10{,}000$ ohms, the result becomes:

$$k = \frac{20 \times 10{,}000}{10{,}000 + 10{,}000} = \frac{200{,}000}{20{,}000} = 10.$$

In this case, when $R_a = R_i$, the amplification of the stage is equal to one-half of μ.

When $R_a = 30{,}000$ ohms, the result becomes:

$$k = \frac{20 \times 30{,}000}{10{,}000 + 30{,}000} = \frac{600{,}000}{40{,}000} = 15.$$

Thus, if $R_a = 3R_i$, the value of k already reaches 75% of the μ value.

And, at last, when $R_a = 90{,}000$ ohms $= 9R_i$, we have the following result:

$$k = \frac{20 \times 90{,}000}{10{,}000 + 90{,}000} = \frac{1{,}800{,}000}{100{,}000} = 18.$$

As may be seen, in this latter case $k = 0.9\mu$.

As R_a increases, the amplification factor of the stage at first grows fast and then slower. Because of this, it is not worth while to employ values of R_a greater than $4R_i$ because too great a value of R_a would cause an excessive d. c. voltage drop and the anode of the valve will not be receiving sufficient voltage necessary for the normal operation of the valve. A noticeably greater gain of the stage will not be obtained by increasing the value of R_a further, because, in any case, the amplification of the stage will always remain smaller than the amplification factor of the valve itself. In practical design of triode amplifier circuits, the value of R_a is made equal to from $3R_i$ to $4R_i$ and then the value of k becomes equal to 75%-80% of the μ value

Every amplifier stage must be so designed that it introduces but the smallest amount of distortion. Some distortion is inevitable and its character must be analysed in each individual design for the purpose of distortion reduction to the minimum. To clarify the nature of distortion, let us refer to a valve characteristic (Fig. 133*a*) and inspect the amplification process in the valve*.

As may be seen from the drawing, alternating sinusoidal grid voltage, having an amplitude (U_{mg}) of 2 volts, is pictured under the characteristic curve. Since grid voltage is laid off along the X-axis, we have to represent the time axis extending downward. The grid voltage curve is shown in an unusual manner—drawn from top to bottom, instead of its common location from left to right.

In the given example, the grid voltage varies from —2 v to +2 v, which corresponds to the linear portion of the curve. Hence, we obtain sinusoidal oscillations of the anode current, the amplitude of such oscillations (I_{ma}) being equal to 3 ma. The curve of the pulsating anode current is plotted to the right of the valve

* In this case, we have to use a dynamic characteristic because valves in amplifier stages work into load resistances.

characteristic, the time axis is horizontal and the current axis—vertical. (In many a drawing, the latter axis is not separately shown, and the one present in the characteristic is utilised for this curve.)

When there is no input signal, i.e., when the grid voltage is zero, the anode current value is 5 ma.

The value of anode current observed when no alternating voltage is applied to the grid is called the resting current and is denoted by I_{a0}. When an amplifier valve operates along the linear part of its

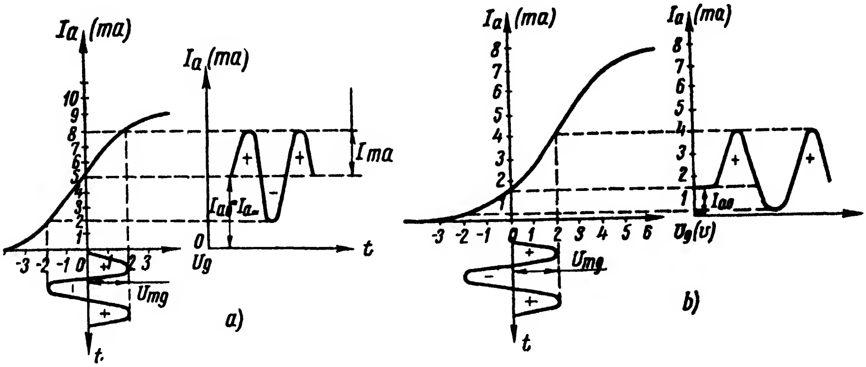


Fig. 133. Principle of amplification of oscillations with the help of a triode

characteristic, the resting current I_{a0} is equal to the d.c. component of the pulsating current obtained under the following condition: $I_{a0} = I_{a-} = 5$ ma.

The point on the curve corresponding to the resting current or, in other words, corresponding to the grid voltage when no alternating signal is applied to the grid, is known as the operating point.

The quality of amplification is determined by the location of the operating point and by the amplitude of alternating current fed to the grid of the valve. If the operating point is located in the middle of the linear portion of the characteristic and if the amplitudes of grid voltage oscillations do not extend beyond this portion, the amplification, practically speaking, is distortionless. Under such condition of operation, the shape of the anode current oscillations corresponds to the shape of the voltage applied to the grid.

Another condition of operation is shown in Fig. 133*b*. Here also the operating point is located on the linear part of the curve but is situated closer to its lower bend. Because of this, the negative half-waves of the voltage being amplified extend to the lower bend and are badly distorted. The amplitude of negative half-waves of the anode current becomes smaller than the amplitude of the positive half-waves, the latter not extending beyond the linear part of

the curve. Non-linear distortion is thus observed when the operation takes place near the bent part of the characteristic.

The section of the characteristic within the limits of which both the grid voltage and the anode current oscillate (i.e., within the limits of which the valve functions) is called the operating section.

Undistorted amplification can be obtained only when the operating section is linear. Such linear section of the valve characteristic lies between the lower bend and the point corresponding to the zero grid voltage. This section is located in the region of negative grid voltages. There is another part of the linear section of the curve — the part extending into the region of positive grid voltages. It is, however, undesirable to use this part of the curve as the operating section of an amplifier, because positive grid voltages give rise to grid current, which is also a cause of non-linear distortion.

Fig. 131 depicts the deleterious effect of grid current. Here, on a negative half-wave of alternating voltage applied to the valve-grid there is no grid current, no load is placed on generator G, and the grid voltage is equal to the generator e.m.f. Grid current, appearing in the valve on a positive half-wave, creates voltage drop in the internal resistance of the generator. In the latter case, the generator begins to work into a load, and the voltage across its terminals (i.e., the voltage applied to the grid) is less than the e.m.f. by a value equal to the voltage drop in the generator.

The internal resistance of such generators is often quite high. For instance, when the generator is represented by a microphone, this resistance is offered by the transformer secondary winding, consisting of a large number of turns and possessing high inductance. As a result, the voltage drop in the generator will also be considerable, and the alternating voltage applied to the grid of the valve will no longer be sinusoidal, i.e., will become distorted. The amplitude of this voltage will be smaller on the positive half-waves than on the negative half-waves.

The higher the amplitude of the alternating grid voltage, the larger will be the grid current and the more pronounced will become the non-linear distortion.

In order to avoid distortion caused by the grid current, the operating section of the characteristic must be located in the region of negative grid voltages. *For this purpose, negative d.c. voltage E_b is applied to the valve grid in an amplifier. This voltage shifts or displaces the operating point to the left and is called the grid bias voltage, or bias voltage, or simply bias.*

Fig. 134*a* illustrates the amplification process in a valve, to the control grid of which the bias voltage is applied. Various methods of generating bias voltage will be explained later.

The amplitude of the alternating voltage applied to the grid of a valve must not be too high, otherwise the oscillations will extend into the region of the lower bend of the characteristic and into the

region of positive grid voltages, where the grid current appears. The value of grid bias voltage E_b and the value of maximum permissible amplitude of alternating voltage $U_{mg\,max}$, fed to the grid of a valve, can be determined by applying a ruler to the curve and locating the linear portion of the characteristic (portion *BC* in Fig. 134*a*). Slight deviations of the characteristic from the straight line are tolerable.

The middle point of the linear portion of the characteristic determines the position of the operating point *A* and bias voltage E_b, while the amplitude value $U_{mg\,max}$ can be considered equal to E_b.

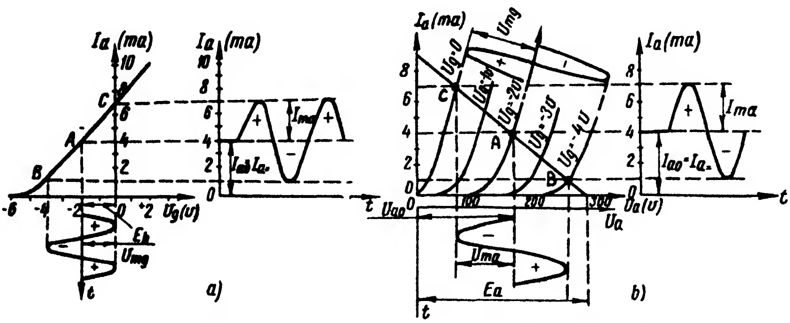


Fig. 134. Amplification of oscillations in the presence of negative grid bias voltage

If the anode voltage is increased, the characteristic will shift to the left and the operating section will be extended. It is possible to amplify large-amplitude signals without distortion by shifting the operating point sufficiently to the left (by increasing the bias voltage).

Valves with the so-called "left-handed" characteristics are the most suitable ones for distortionless amplification. In such valves, the characteristic is located in the region of negative grid voltages, has a longer operating section than a "right-handed" valve, and amplifies even large-amplitude signals without distortion.

The location of the operating point is determined by the anode voltage and bias. These two values determine the *operating condition of the valve*. Filament or heater voltage, of course, also affects the position of the operating point but it is assumed that this voltage always has a normal, rated, value.

Anode characteristics are often resorted to for the purpose of graphic representation and calculation of the amplification process. This is shown in Fig. 134*b*, dealing with a similar case as Fig. 134*a*. Plotting of anode dynamic characteristics was discussed in Sec. 36. A selected operating point *A* determines the value of bias voltage (—2 v), while the amplitude of voltage being amplified (U_{mg}) deter-

mines the operating section BC. After the operating section has been selected, it becomes easy to plot the curves of anode current and anode voltage changes, as shown in Fig. 134*b*. At the same time, the anode current curve also shows voltage changes across the load resistance connected into the anode circuit.

The operating point and the value of load resistance R_a, upon which depends the slope of the dynamic characteristic, should be so selected that sections BA and AC are made as equal as possible. Such procedure assures practically undistorted amplification.

Let us now consider the question of power in an amplifier stage. When no signal is applied to the grid, a steady direct current flows through the anode circuit. The power, which is a function of this current, is partly consumed in resistance R_a and partly dissipated by the anode of the valve. When alternating voltage is applied to the valve grid, the anode current begins to pulsate and its a.c. component develops *useful power* $P_\sim$ in resistance R_a, the level of this power being determined by one of the following formulae:

$$P_\sim = \tfrac{1}{2} I_{ma}^2 R_a = \tfrac{1}{2} I_{ma} U_{mR} = \tfrac{1}{2} \frac{U_{mR}^2}{R_a}.$$

The coefficient $\tfrac{1}{2}$ has been introduced into these formulae, because we deal with amplitude values of current and voltage. The *input power* (the d.c. power consumed by the anode circuit of an amplifier stage) is calculated as

$$P = I_{a=} E_a,$$

i.e., as a product of d.c. component of anode current $I_{a=}$ and voltage E_a of the anode power supply.

The ratio of the useful power $P_\sim$ to the input power P is called the *efficiency* of the stage. Denoting the efficiency by Greek letter η (eta), the efficiency is given by the following expression.

$$\eta = \frac{P_\sim}{P}.$$

Voltage amplifiers do not require high efficiency because they handle insignificantly small power. Accordingly, the efficiency of voltage amplifier stages does not exceed several per cent. On the other hand, the efficiency of output stages, handling considerable power, is required to be dozens of per cent.

Considerations of energy economy call for the operation of valves under the conditions of minimum input power. This is obtained by the reduction of the $I_{a=}$ value.

The latter is achieved by shifting the operating point to the left as far as possible. Hence, the correct choice of bias voltage not only serves to reduce distortion but also to secure the most economical operating condition (the highest efficiency) of an amplifier. It is, of course, undesirable to apply too large a negative bias voltage,

as this can bring the operating section down to the lower bend, reduce the a.c. component of the anode current (i.e., reduce the amplification), and cause non-linear distortion.

The efficiency of an amplifier stage should not be confused with the *coefficient of power amplification (gain), k_p, of the stage.* Gain k_p is the ratio of the useful power $P_\sim$ to grid power P_g, expended by generator G in the grid circuit:

$$k_p = \frac{P_\sim}{P_g}.$$

In low-frequency amplifiers, k_p is not as important as the voltage amplification factor k.

If a stage operates without grid current, grid power P_g is insignificantly small. In this case, P_g is determined only by leakage currents arising due to imperfections of insulation in the grid circuit, and is also determined by capacitance current flowing through the input circuit of the stage (between the grid and cathode), this current accounting for a certain loss of power in the internal resistance of generator G. However, with normal insulation, the leakage current is quite insignificant; and the capacitance current on low frequencies is equally insignificant as the input capacitance is very low (a few dozens of picofarads) and its capacitive reactance is usually very high.

Because of very small value of P_g, low-frequency amplifiers working without grid current are noted for very high values of k_p. On the other hand, in an amplifier operating under conditions of grid current, the value of k_p is sharply reduced. A reduction of k_p also takes place with an increase of frequency owing to the increased energy loss in the grid circuit as the frequency is raised. On ultra-short waves, for instance in the decimetric wave band, the value of k_p may be lowered to 1, making the use of an amplifier stage pointless.

72. RESISTANCE-COUPLED AMPLIFIERS

Resistance-coupled amplifiers represent the most popular variety of low-frequency amplifiers. In this variety, the load resistance R_a is, actually, an ohmic resistor. Fig. 135 shows a single-stage resistance-

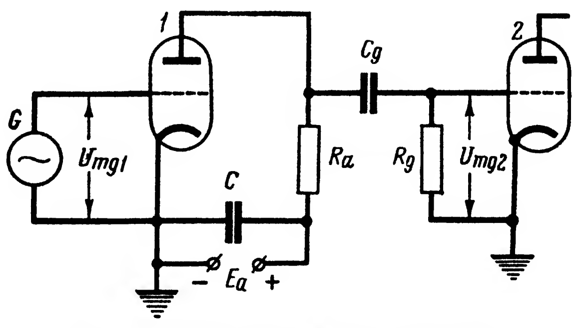


Fig. 135. Resistance-coupled amplifier

coupled amplifier and the connection of this stage with the valve of the following stage.

Alternating voltage U_{mg1}, supplied by generator G, is fed to the grid of the first valve for the purpose of amplification. The anode

current of the first valve becomes pulsating, its d.c. component flowing through the load resistor R_a and anode-current-power supply, while its a.c. component passes through the load resistor and the capacitor C. The passage of the a.c. component of the anode current through resistor R_a builds up an amplified alternating voltage across this resistor.

A resistance-coupled stage generally feeds the amplified voltage to the grid of the next valve, i.e., serves as a voltage amplifier. It is, however, impossible to feed this voltage directly to the grid-cathode circuit of the driven stage because this would apply high positive voltage from resistor R_a to the control grid of valve *2*. This is why the alternating voltage taken off across R_a is fed to the next stage through *grid capacitor* C_g. This capacitor insulates the grid of valve *2* from the high d.c. voltage of the anode power supply, while freely passing the alternating current signal.

The coupling capacitor is always used in conjunction with *grid resistor* R_g — otherwise called the *leakage resistor*. If resistor R_g were not provided, the electrons reaching the control grid would be accumulating on it. In a short period of time, the negative grid potential would be increased to such an extent that the anode current of valve *2* would be cut off and the valve would stop functioning. When the resistor is connected into the circuit as shown in Fig. 135, it "drains" the electrons off the grid; thus a current flows in the grid circuit and the accumulation of electrons on the grid is prevented.

Thus, the alternating potential from the anode end of resistor R_a (the upper end in the drawing) is fed through capacitor C_g to the grid of valve *2*, the opposite polarity of such alternating potential being fed from the other (lower) end of R_a to the cathode of valve *2* through capacitor C and the "common negative" circuit. Let us denote the voltage fed to the grid of valve *2* as U_{mg2}. The amplification factor k of the amplifier stage is then expressed as:

$$k = \frac{U_{mg2}}{U_{mg1}}.$$

When studying the operation of the amplifier stage based on valve *1*, it is customary to consider components R_g and C_g as a part of the load of this valve, the resistor and capacitor acting as an additional load to the main load represented by resistor R_a. Thus an amplifier stage is considered as beginning with the grid of valve *1* and ending with the grid of valve *2*. However, when studying the operation of the amplifier stage based on valve *2*, we have to consider components C_g and R_g as they are connected into the grid circuit of this valve. These components couple the first stage to the second and, therefore, can be thought of as belonging to either one of the two stages.

When a valve functions under the conditions of dynamic operation, its anode voltage changes are always in phase opposition to grid voltage changes (see Fig. 134*b*). Hence, the amplified alternating

voltage is always in phase opposition to the alternating voltage applied to the grid. It is customary to say that an amplifier stage "inverts" the voltage phase, referring the voltage, of course, to the potential of the cathode (common negative or earth).

In resistance-coupled triode amplifiers, the value of R_a ranges within $3R_i$ and $4R_i$ and amounts to several tens of thousands or hundreds of thousands of ohms. The capacitance of capacitor C_g is of the order of 5,000-100,000 pf. This means that its capacitive reactance is quite small even on low frequencies. The value of resistor R_g is usually several times higher than the value of resistor R_a, i.e., is generally equal to anything from 0.1 to 1 megohm. It is not advantageous to use low values of R_g, because the resistor is connected in parallel with R_a through capacitors C_g and C. Because of this, too low a value of R_g would decrease the value of load resistance of valve *1*, thus lowering the amplification factor of the stage. On the other hand, too high a value of R_g is also undesirable, because in this case the electrons would not leak off the grid of valve *2* in time. As a result, the anode current of the valve would at times be cut off by the electrons accumulated on the grid, particularly on strong signals.

If the operating condition of a valve used in a resistance-coupled amplifier has been correctly selected and the operating point is located on the linear part of characteristic, non-linear distortion of the stage will be very small. Such amplifier, employed as a voltage amplifier, usually operates on small a.c. voltages applied to its grid and, because of this, will produce only insignificant distortion. Negative grid bias voltage, fed to this type of amplifier, prevents distortion-causing grid current and also reduces the consumption of power by the anode circuit (see Sec. 75).

For a considerable band of frequencies lying in the middle-frequency range, it is permissible to neglect the effect of capacitances and to consider the load resistance of an amplifier stage as a pure ohmic resistance consisting of R_a and R_g connected in parallel. Fig. 136*a* gives an equivalent circuit for the middle frequency band. The frequency characteristic is ideal for these frequencies and there is no frequency distortion.

On the lower and higher parts of the sound-frequency range the amplification is reduced and attenuation becomes noticeable. Hence, the frequency characteristic has an appearance as in Fig. 129 *a* and *b*.

The reduction of amplification on the lower frequencies is due to the influence of the coupling capacitor C_g. As the frequency is lowered, the capacitive reactance of this capacitor increases, the voltage drop across it increases, and the voltage applied to the grid of valve *2* is reduced. An equivalent circuit for the lower frequencies is given in Fig. 136*b*. Capacitor C is not shown in this circuit because this capacitor has a large value of capacitance and its capacitive

reactance is negligible even on low-frequency currents (for example, the filtering capacitor of a rectifier has a value of several microfarads).

On higher audio frequencies, the capacitive reactance of C_g becomes very low and can be disregarded. However, on these higher frequencies the input capacitance C_i of the valve (i.e., the capacitance between the grid and cathode of valve *2*) begins to tell. This capacitance is also supplemented by the output capacitance, i.e.,

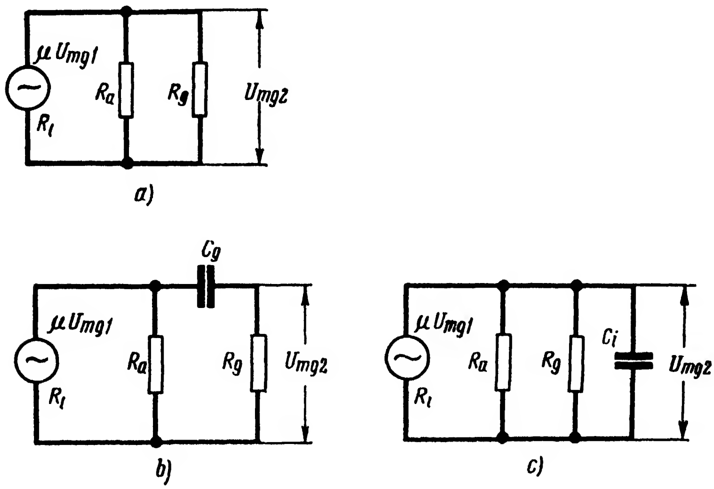


Fig. 136. Equivalent circuits of a resistance-coupled amplifier: *a)* for medium frequencies; *b)* for lower frequencies; *c)* for higher frequencies

by the capacitance existing between the anode and cathode of valve *1* and by the capacitance of the wiring. These capacitances are, however, considerably smaller than C_i and can be disregarded.

An equivalent circuit for this case is shown in Fig. 136*c*. The value of capacitance C_i is comparatively small and does not exceed some tens of picofarads. On the lower and middle frequencies it has no influence upon the amplification because its capacitive reactance is very high. However, on the higher audio frequencies, the capacitive reactance of C_i is reduced and the total impedance of the parallel-connected R_a, R_g and C_i is reduced accordingly. This reduces the amplification factor of the stage.

The attenuation on the extreme frequencies of the audio range is comparatively small. Hence, the main advantage of a resistance-coupled amplifier is seen in that it creates insignificant frequency distortion. Its another advantage is the circuit simplicity. But the voltage drop across resistor R_a in the anode circuit is the disadvantage of this type of amplifier. Due to this shortcoming, the voltage

U_a applied to the anode of a resistance-coupled amplifier is always much lower than the anode power-supply voltage E_a. This is clear from the following relation:

$$U_a = E_a - I_{a=}R_a.$$

Resistance-coupled amplifiers employ low-power triodes and high-frequency pentodes, the latter also giving good performance in low-frequency circuits. Pentode valves, of course, cannot employ values of R_a as high as $3R_i$ or $4R_i$, since in these valves the value of R_i may be as high as several million ohms. A few hundreds of thousands of ohms is the usual value of R_a in pentode circuits. In usual cases, R_a equals from $0.05R_i$ to $0.2R_i$. The amplification factor of the stage (k) is considerably lower than μ. However, since pentode valves possess very high values of μ, they give higher amplification than triodes, even when k is only 5-20% of μ.

Example. An amplifier stage employs the triode part of valve, type 6Г7, with the following parameters: $\mu = 70$; $R_i = 60{,}000$ ohms. The load resistor connected into the anode circuit of the valve has been computed as follows: $R_a = 4R_i = 4 \times 60{,}000 = 240{,}000$ ohms. Then, the amplification factor of the stage is given by the following equation:

$$k = \frac{70 \times 240{,}000}{60{,}000 + 240{,}000} \approx 56.$$

Now, assume that the same amplifier employs a 6Ж7 pentode instead of the 6Г7 triode. The parameters become: $\mu = 1{,}400$; $R_i = 1{,}200{,}000$ ohms. The value of R_a is left unchanged and, hence, $R_a = 0.2R_i$. The amplification factor of the stage becomes:

$$k = \frac{1{,}400 \times 240{,}000}{1{,}200{,}000 + 240{,}000} \approx 230.$$

Although in the case of the pentode, R_a is equal only to 20% of R_i, the amplification has increased four times because of the greater amplification factor of pentode.

The necessity of employing for pentodes definite values of load resistors, considerably smaller than R_i, follows from the study of anode characteristics of the valves. Fig. 137, related to a pentode operating in a resistance-coupled amplifier, gives the anode characteristics for three different values of load resistance. The bias voltage and the amplitude of alternating voltage applied to the grid are the same for all the three cases. The value of resistance R_{a1} is too low. The dynamic characteristic relating to it is too steep and

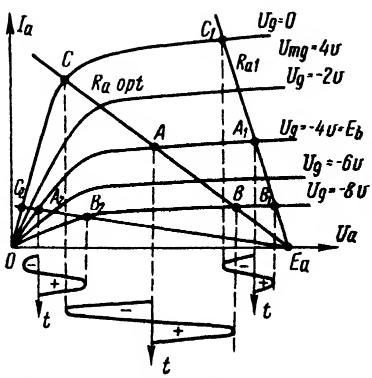


Fig. 137. The selection of an optimum operating condition for a pentode

the amplification of voltage is comparatively small. Besides this, portions A_1C_1 and A_1B_1 of the operating section, corresponding to the negative and positive half-waves of the voltage being amplified, are not equal, and, consequently, give rise to considerable non-linear distortion.

A still less favourable case is observed when the load resistance is excessive and when the dynamic characteristic is nearly level.

The operating section of the characteristic is short. The amplification of voltage is small, while the non-linear distortion is considerable, because in this case the portion of operating section A_2C_2 for the negative half-wave of the voltage under amplification is larger than portion A_2B_2. For a certain intermediate value of load resistance, i.e., for an optimum, the most advantageous value, ($R_{a\ opt}$), the dynamic characteristic passes in such a way that portions AC and AB are equal and the non-linear distortion is small. Under such operating conditions, the value of amplified voltage becomes considerable. This optimum operating condition is not difficult to select with the help of a ruler.

Drawing the dynamic characteristic for such an operating condition, the value $R_{a\ opt}$ is found by the division of E_a by the current corresponding to the point of intersection of the dynamic characteristic with the Y-axis. Owing to the fact that the slope of the most advantageous dynamic characteristic for $R_{a\ opt}$ is considerably smaller than the slopes of static characteristics of pentodes, the value of $R_{a\ opt}$ is several times smaller than R_i.

In the pentode-amplifier stage, R_a is many times smaller than R_i; therefore, in the basic formula for coefficient of amplification

$$k = \frac{\mu R_a}{R_i + R_a}$$

the value of R_a in the denominator can be ignored in the presence of R_i.

Thus, we have

$$k \approx \frac{\mu R_a}{R_i},$$

but

$$\frac{\mu}{R_i} = S,$$

hence the approximate value of coefficient of amplification of a pentode stage can be found by the simpler formula:

$$k = SR_a.$$

Let us now briefly consider the question of components. Resistance R_a must be designed for the power to be dissipated by the given resistor. For instance, if the d.c. component of the anode current is given by $I_{a-} = 5$ ma, and $R_a = 20{,}000$ ohms, then the power is equal to $P = I^2_{a-}R_a = 0.005^2 \times 20{,}000 = 0.5$ watt. The resistance, then, must be designed for this power. Usually non-wire-

wound resistors are used in such power ratings. The grid resistor R_g can be designed for the smallest power dissipation because the grid current it carries has a negligible value. Capacitor C_g must have good insulation. If this capacitor is leaky, it will pass high positive voltage from a preceding stage to the grid circuit of the next stage. It is undesirable to use a high-capacitance paper capacitor as the coupling capacitor (C_g), because capacitors of this type, as a rule, have rather poor insulation. At the same time, the capacitance of C_g should be high to prevent a decrease of amplification on lower frequencies. Besides, the capacitor used as C_g must be able to withstand high anode voltage and, hence, its voltage rating should be correspondingly high. The best type of capacitor for this use is a mica capacitor having a capacitance value of several tens of thousands of picofarads.

73. CHOKE-COUPLED AMPLIFIERS

A choke-coupled amplifier (Fig. 138) differs from a resistance-coupled amplifier in that it employs an iron-core low-frequency choke L instead of the ohmic resistor R_a. To obtain a high value of inductance, the choke is wound with approximately 10,000-20,000 turns of wire. This gives the choke an inductance of several tens of henries and, consequently, high inductive reactance, which is necessary if we are to obtain a high value of amplification in a choke-coupled amplifier where the choke serves as the load connected into the anode circuit.

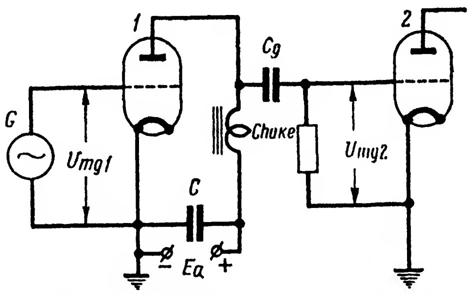


Fig. 138. Choke-coupled amplifier

The general operating principle of a choke-coupled amplifier is similar to that of a resistance-coupled amplifier. The resistance of the choke winding is comparatively small (not over several hundreds of ohms) and, therefore, there is but a small voltage drop in the anode load. This is the advantage of a choke-coupled amplifier over amplifiers using resistance coupling; and it can be assumed, in practical cases, that in choke-coupled amplifiers the voltage actually fed to the anode (U_a) is the same as the voltage developed by the anode power supply (E_a). The amplification factor of a choke-coupled stage is always less than μ, just as it is in the case of a resistance-coupled stage.

A choke-coupled amplifier introduces a greater frequency distortion than does a resistance-coupled amplifier, and this is the

disadvantage of amplifiers employing the choke-coupling principle. In a choke-coupled amplifier, the frequency characteristic is more attenuated on higher and lower frequencies. On the lower frequencies, the amplification is decreased not only because of the increased capacitive reactance of capacitor C_g but also because of the decreased inductive reactance of the coupling choke L.

On higher audio frequencies, the inductive reactance x_L of the choke is very high, but it is shunted by two capacitances, one of which is the input capacitance C_i of the following valve (this, of course, also holds true for a resistance-coupled amplifier, as already explained above). The other one of the undesirable but inevitable capacitances in a choke-coupled amplifier is the distributed capacitance of the windings of the choke. Such distributed capacitance can be as high as 100-200 pf; on the higher audio frequencies the total capacitive reactance of this capacitance and of C_i is comparatively small. As a result amplification at these frequencies decreases. For instance, if the stage employs a valve with parameters $\mu = 100$ and $R_i = 60{,}000$ ohms, and if the inductance of the choke is equal to 40 h, then the inductive reactance for the middle frequency, 400 cps, will be given by the following:

$$x_L = 6.28\, fL = 6.28 \times 400 \times 40 \approx 100{,}000 \text{ ohms},$$

i.e., x_L is greater than R_i and the amplification of the stage will be sufficient. At the same time, when the amplifier operates on the frequency of 100 cps, the value of x_L will be 4 times smaller and equal only to 25,000 ohms. In this case, x_L is smaller than R_i, and the amplification will be considerably decreased.

Assume that the choke has a distributed capacitance (C_L) of 200 pf, and the input capacitance C_i of the next stage is 150 pf. On one of the higher frequencies, say on 8,000 cps, the inductive reactance of the choke will be very high; it will be 20 times as great as on 400 cps, i.e.,

$$x_L = 20 \times 100{,}000 = 2{,}000{,}000 \text{ ohms}.$$

This, however, does not alleviate matters, because on the given frequency the capacitive reactance of the combined shunting capacitance (C), made up of C_i and C_L, has a low value, seen from the following:

$$x_C = \frac{1}{6{,}28 fC} = \frac{10^{12}}{6{,}28 \times 8{,}000 \times 350} = 56{,}000 \text{ ohms}.$$

This value is less than the value of x_L on 400 cps, and, consequently, the amplification will be reduced on the higher frequencies as well.

In choke-coupled amplifiers specially designed chokes have to be used if the frequency distortion is to be kept to a reasonable value. A choke of this type is provided with an air gap to prevent

the inductance of the choke from decreasing. Such inductance decrease can be caused by the magnetic saturation of the core by the d.c. component of the anode current. Besides this precaution, the winding of the choke must be sectionalised to reduce its distributed capacitance.

Choke-coupled amplifiers are considerably less popular than resistance-coupled amplifiers.

74. TRANSFORMER-COUPLED AMPLIFIERS

Transformer-coupled amplifiers (Fig. 139) are quite frequently employed in various radio equipment. In this type of amplifier, the primary winding of interstage transformer T is connected into the anode circuit of valve *1* and serves as the anode load. The secondary winding of the transformer feeds the signal voltage to the grid circuit of the following stage. The ratio of transformation, i.e., the ratio of number of turns of primary winding w_1 to the number of turns of the secondary winding w_2 is denoted as follows:

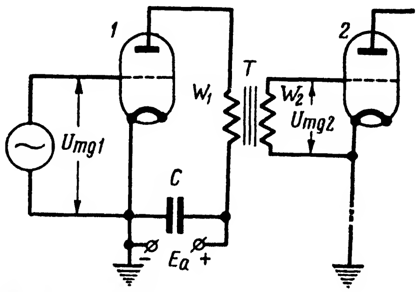


Fig. 139. Transformer-coupled amplifier

$$n = \frac{w_1}{w_2}.$$

The value of n in interstage transformers can be anything from 1 : 1 to 1 : 4. The primary winding usually contains from 3,000 to 6,000 turns of wire. Both the primary and the secondary windings employ 0.08-0.12 mm enamelled wire.

The cross-sectional area of the transformer core may be from 1 to 6 sq cm. The primary winding has a low resistance, and it may be assumed that $U_a = E_a$.

High gain of a transformer-coupled amplifier stage is the advantage of this type of coupling. When a step-up interstage transformer is employed, this factor can be higher than the μ of the valve, which is impossible to obtain with other types of interstage coupling. Another advantage of a transformer-coupled amplifier is seen in good isolation of the grid circuit of the driven stage from the anode circuit of the driving (i.e., preceding) stage; the two windings of the transformer are well insulated from each other and there is no coupling capacitor to create current leakage. It is because of this isolating property that some amplifiers resort to transformer coupling, employing interstage tranformers with a radio of 1 : 1, although such transformers provide no voltage gain.

The ability of a transformer to convert not only voltage and current but also the value of load impedance is an important feature of transformer-coupled circuits.

Let us illustrate this feature with the following example.

Assume that the transformation ratio n of a certain step-down transformer is equal to 4 : 1. Also assume, that the value of the load resistance R_2 is equal to 20 ohms, while the generator voltage U_1 (i.e., the voltage of the primary winding) is equal to 160 v (Fig. 140*a*).

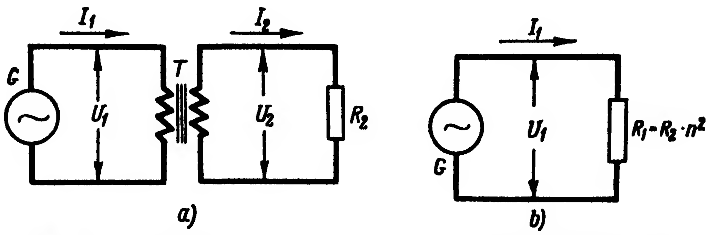


Fig. 140. Connection of a load resistor through a transformer and an equivalent circuit

From the above-stipulated conditions, we have the following:

secondary winding voltage $U_2 = U_1 : n = \frac{160}{4} = 40$ v;

secondary circuit current $I_2 = U_2 : R_2 = \frac{40}{20} = 2$ a;

secondary circuit power $P_2 = U_2 I_2 = 40 \times 2 = 80$ w.

Since the power losses in a transformer are usually very small, it can be considered that the primary circuit power P_1 is also equal to 80 w. Hence: $I_1 = P_1 : U_1 = \frac{80}{160} = 0.5$ a.

It now becomes possible to calculate resistance R_1, which is the resistance offered by the primary winding of the transformer to the generator.

$$R_1 = U_1 : I_1 = \frac{160}{0.5} = 320 \text{ ohms}.$$

As seen from above, R_i is 16 times greater than R_2. But 16 is equal to 4^2, i.e., the square of transformation ratio n. Hence, the following relation may be written:

$$R_1 = R_2 n^2.$$

We have arrived at the following important conclusion: ***a transformer loaded with a certain resistance R_2 offers resistance R_1 to the generator, the latter resistance being n^2 times greater than resistance R_2.*** In other words, a transformer possessing a transformation ratio of n converts the value of the load resistance n^2 times. If the gener-

ator were connected directly to resistance R_2, it would be working into a 20-ohm load resistance. However, since the load R_2 is actually connected to the generator through a transformer with $n = 4$, the generator works into load resistance R_1, which is equal to $20 \times 4^2 = 320$ ohms.

Thus, a step-down transformer increases the load resistance offered to the generator. If R_2 is increased, R_1 will also increase proportionally. A decrease of R_2 will decrease R_1. However, with all these changes, R_1 will remain greater than R_2 by n^2 times. When $R_2 = 0$, i.e., when the secondary circuit is short-circuited, R_1 will also be equal to zero. This creates a condition of short circuit in the primary circuit as well, the transformer will draw an abnormally heavy current from the generator, and the latter will, thus, also be short-circuited. In fact, in the case just described, only the resistance of the secondary winding of the transformer will remain in the circuit, while the primary circuit will consist only of the internal resistance of the generator and of the resistance of the primary winding.

The latter two resistances will limit the current increase under the condition when the transformer is short-circuited. When the secondary winding is open (the transformer operates without load), i.e., when R_2 value may be considered very high, R_1 value is also very large and transformer will draw very small current from the generator (no-load current). It may be considered in this case that the primary circuit is also open, so to speak. Under such conditions, the transformer acts simply as a choke with high inductive reactance, because the secondary winding is not actually functioning.

Dealing with a step-up transformer, it can be shown that it, too, converts the resistance by n^2 times. In this case, however, a reduction of the load resistance is obtained.

Let us take, for example, a step-up transformer with the following characteristics: $n = 1:3$; $U_1 = 90$ v; $R_2 = 2{,}700$ ohms. Now we calculate the remaining values: $U_2 = 90 \times 3 = 270$ v; $I_2 = \frac{270}{2.700} = 0.1$ a; $P_2 = 270 \times 0.1 = 27$ w.

Considering that $P_1 = P_2 = 27$ w, $I_1 = \frac{27}{90} = 0.3$ a, $R_1 = \frac{90}{0.3} = 300$ ohms, it becomes clear that R_1 is 9 times smaller than R_2. But in this case, $n = \frac{1}{3}$ and the following relation can be written:

$$R_1 = 2{,}700 \left(\frac{1}{3}\right)^2 = 2{,}700 \times \frac{1}{9} = \frac{2{,}700}{9} = 300 \text{ ohms.}$$

As follows from above, the same resistance conversion law $R_1 = R_2 n^2$ also holds true for a step-up transformer.

Because resistance R_1 depends upon the value of R_2, it is customary to say that R_1 is the resistance of the load ***related to the primary winding or transposed to the primary circuit.***

It is possible to picture an equivalent circuit of the primary circuit of transformer as shown in Fig. 140*b*. In other words, it can be considered that the generator works into load R_1, this resistor taking place of a transformer, whose secondary winding is loaded with resistance R_2.

The aforesaid conversion of resistance (or impedance) with the help of a transformer is of great significance in the operation of a transformer-coupled amplifier stage, and is of particular importance in output stages.

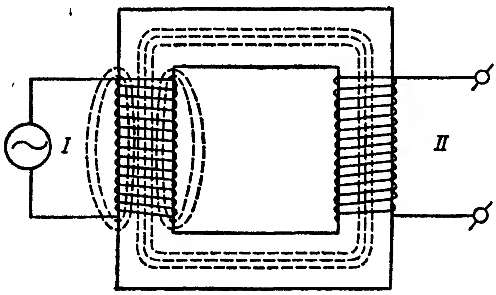


Fig. 141. The effect of magnetic leakage in a transformer

Every transformer possesses some magnetic leakage, which is explained as follows. The whole of the magnetic flux set up by the primary does not interlink with the turns of the secondary winding. A certain part of the flux passes only around the primary winding and does not participate in the creation of the secondary-winding e.m.f. (Fig. 141). It may be said that some of the primary winding turns do not participate in the process of voltage and current transformation, these turns only constituting a certain inductance connected in series with the working turns of the primary winding. This inductance is called the leakage inductance and is denoted by L_{l1}.

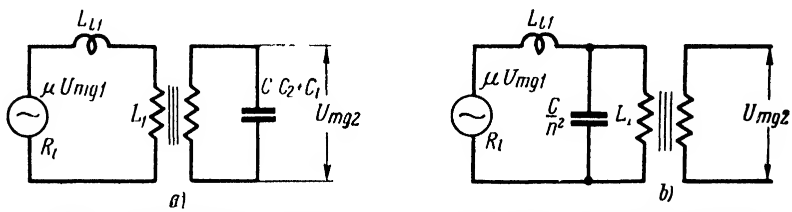


Fig. 142. Equivalent circuits of a transformer-coupled amplifier

Fig. 142*a* shows a simplified equivalent circuit of an amplifier stage, where the valve is represented as a generator. This circuit also includes leakage inductance L_{l1} and inductance L_1 of the working turns of the primary winding. In the given circuit, the secondary circuit of the transformer includes capacitance C, the latter made up of the distributed interturn capacitance C_2 of the secondary winding and supplemented by the input capacitance C_i of the following stage.

The circuit does not show the distributed capacitance of the primary winding, because this capacitance does not play a significant role in the operation of the circuit. Thus, the interstage transformer is loaded with the capacitive reactance x_C of the parallel-connected capacitances C_2 and C_i. This capacitive reactance may be regarded in relation to the primary winding, i.e., it may be considered as connected into the primary circuit, if the given reactance is multiplied by n^2. In other words:

$$x_C n^2 = \frac{1}{6{,}28 f C} n^2.$$

This expression may be transposed in the following way:

$$x_C n^2 = \frac{1}{6{,}28 f \frac{C}{n^2}}.$$

Apparently, loading of the secondary circuit with capacitance C is equivalent to connecting capacitance $\frac{C}{n^2}$ into the primary circuit of the transformer. In accordance with this, Fig. 142*b* gives an equivalent circuit of a transformer-coupled amplifier stage in which capacitance C is transposed into the primary circuit of the transformer.

Referring to this circuit, it becomes possible to analyse the behaviour of a transformer-coupled amplifier on different audio frequencies and to determine the resulting character of frequency distortion.

As far as the lower audio frequencies are concerned, the inductive reactance of L_{l1} is very small and need not be taken into consideration. It is also permissible to disregard the capacitive reactance of the shunting capacitance $\frac{C}{n^2}$, because it is many times as great as the inductive reactance L_1 of the working turns. Thus, only the inductance is of importance on the lower frequencies and, in this case, the equivalent circuit is composed as shown in Fig. 143*a*. However, on the lower audio frequencies the inductive reactance, $x_L = 6.28 = fL_1$, is comparatively low and, hence, the amplification factor of the stage will also be low. Here, a decrease of amplification of the lower frequencies will be experienced, just as it was experienced with the other types of amplifiers.

To illustrate this, let us take a numerical example.

Let $L_1 = 40$ henries. L_{l1}, which is usually about 1% of L_1, then becomes equal approximately to 0.4 h. The remaining constants are: $C_2 = 200$ pf; $C_i = 50$ pf; $n = 1:2$. The capacitance C of the secondary circuit amounts to $C = 250$ pf. Relating this capacitance to the primary winding, we have:

$$\frac{C}{n^2} = \frac{250}{(1/2)^2} = 250 \times 4 = 1{,}000 \text{ pf}.$$

On 50 cps, the inductive reactance of the working turns is given as $x_L = 6.28 \times 50 \times 40 \approx 12{,}500$ ohms. The inductive reactance of the leakage turns, taken as 1% of this value, is equal to 125 ohms. The capacitive reactance shunting L_1 is given as:

$$x_C = \frac{10^{12}}{6.28 \times 50 \times 1{,}000} \approx 3{,}200{,}000 \text{ ohms.}$$

It is apparent that we can neglect the influence of the reactance of inductance L_{l1}. The reactance of the shunting capacitance can be also disregarded.

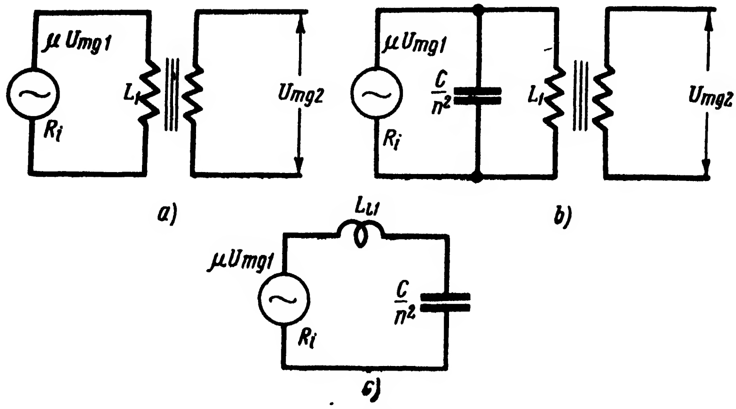


Fig. 143. Equivalent circuits of a transformer-coupled amplifier: *a)* for lower frequencies; *b)* for medium frequencies; *c)* for higher frequencies

Coming to the medium audio frequencies, we can continue to disregard the influence of L_{l1}, because, in the case being described, the reactance of this constant is equal only to 1% of reactance of the inductance L_1. The load reactance, given by $x_L = 6.28fL_1$, will be greater and, in accordance with this, the amplification factor of the stage will also be increased. On a certain medium frequency, the reactance of inductance L_1 and the reactance of the shunting capacitance $\frac{C}{n^2}$ can become equal to each other, which is likely to result in parallel resonance. Under condition of such resonance, as we already know from the previous studies, the impedance of the tuned circuit is at the maximum value. This will increase the amplification factor of the stage on medium audio frequencies. The equivalent circuit for such frequencies is shown in Fig. 143*b*.

Regarding the higher audio frequencies, we note that the reactance of L_1 is many times greater than the reactance of $\frac{C}{n^2}$ and, therefore, L_1 need not be taken into consideration. The equivalent circuit for this condition is shown in Fig. 143*c*.

The reactances of the remaining components L_{l1} and of $\frac{C}{n^2}$ may become equal at a certain frequency, giving rise to a series resonance. As a result, a considerable alternating voltage can develop across capacitance $\frac{C}{n^2}$ and, hence, at the grid of the following valve. In the example given above, a frequency of 8,000 cps gives the following inductive reactance for x_L: $x_L \approx 6.28 \times 8{,}000 \times 0.4 \approx$ $\approx 20{,}000$ ohms, while the reactance of the capacitance becomes:

$$x_C = \frac{10^{12}}{6.28 \times 8{,}000 \times 1{,}000} \approx 20{,}000 \text{ ohms},$$

which indicates that a series resonance will appear on 8,000 cps and the amplification will be sharply boosted up.

The frequency characteristic of a transformer-coupled amplifier stage is shown in Fig. 144. It is apparent that such a stage produces greater frequency distortion, when compared to resistance-coupled and choke-coupled stages. Moreover, the amplification of a transformer-coupled stage is unnaturally increased at a certain high audio frequency, after which, as the frequency is further increased and the resonance point is passed, a sharp decrease of amplification follows.

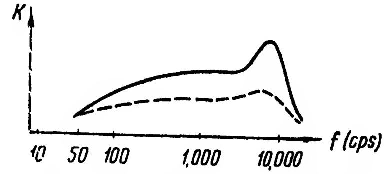


Fig. 144. Frequency characteristics of a transformer-coupled amplifier

The latter effect is attributed to the increase of voltage-loss in the leakage inductance and also to the decrease of the shunting capacitive reactance. It should be noted, that the series resonance in the output transformer-coupled stage usually occurs not at one of the higher audio frequencies, but at a still higher frequency. This is explained by the fact that the output stage has no next stage to drive and, therefore, no input capacitance C_i is encountered by it. The resonance frequency of the output stage is raised also because the distributed capacity of windings, belonging to this stage, is very small.

When the inductance of primary winding L_1 is sufficiently high, the reduction of amplification on the low frequencies will not be very pronounced. It is practically impossible, however, to wind the primary winding with a very large number of turns, as the dimensions of the transformer would become prohibitively large and the interturn distributed capacitance would be too high. This is the reason why the primary winding of an average interstage transformer cannot be designed for larger inductive reactance than a few thousands of ohms on the lower audio frequencies.

To obtain a reasonable degree of low-frequency amplification with such low reactance, it becomes necessary to use valves with low anode resistance R_i in transformer-coupled amplifiers. Triode

valves designed for voltage-amplification duty are manufactured in the following two varieties: 1) triodes with high μ and high R_i — for use in resistance-coupled amplifiers; 2) triodes with average values of μ and R_i — for use in transformer-coupled amplifiers.

However, as far as power triodes designed for output stages are concerned, they always have a small value of R_i and are quite suitable for use in transformer coupled circuits.

As shown above, the resonance phenomenon occurs on the higher frequencies. The best method of counteracting such resonance and its detrimental effects is to make the resonance frequency so high that it would fall above the audio frequency range handled by the amplifier, i.e., at some frequency beyond audibility. To do this, it becomes mandatory to decrease the leakage inductance L_{l1} and the distributed capacity of transformer windings. This is achieved by using sectionalised windings and by decreasing the insulation thickness between the primary and the secondary, thus placing the windings closer to each other. Another, and a simpler, method is that of increasing the damping of the circuit, consisting of L_{l1} and $\frac{C}{n^2}$, by loading the secondary winding with an ohmic resistance having a value of several tens of thousands or hundreds of thousands of ohms.

Such a resistance is connected in parallel with capacitance C and is known as the secondary winding shunting resistor. This resistor damps the circuit and dulls the resonance.

The broken line in Fig. 144 shows the frequency characteristic of a transformer-coupled amplifier in which such shunting resistor is connected across the secondary winding. In the same drawing, the solid line shows the characteristic of the same amplifier using no such resistor. The lower the value of the shunting resistor, the greater is the damping of the circuit and the better is dulled the resonance. This is attributed to the fact that as the resistance value is lowered more and more, the larger part of the current will be passing through it instead of passing through the capacitance. Since the resonant effect pertains only to the part of the current flowing through the capacitance, the resonance becomes less and less pronounced with the decrease of the shunting resistor value.

The disadvantage of the described method of resonance-suppression is seen in that such a system decreases the amplification of an amplifier not only on the resonance frequency but on all the other frequencies as well. An auxiliary short-circuited winding is sometimes used instead of the shunting resistor. Such a winding, consisting of several turns, damps the circuit and dulls the resonance, but also decreases the amplification on all the other frequencies, acting in a manner similar to the shunting resistor.

A transformer is also a source of non-linear distortion. The reason for this is the magnetic saturation of the transformer-core under the influence of d.c. anode current flowing through the transformer winding. When the core is saturated, the changes of its magnetic flux will no longer be proportional to current changes in the primary winding. Such a case is shown in Fig. 145*a*. This drawing gives the core magnetisation curve, i.e., the curve showing the dependence of magnetic flux Φ upon current I_a flowing through the primary winding. In the case of a high value of d.c. component $I_{a=}$ of the anode current, the transformer functions in the region close to magnetic saturation and, therefore, the a.c. component $I_{a\sim}$ creates distorted changes of magnetic flux $\Phi_{\sim}$.

The alternating voltage set up in the secondary winding by this magnetic flux will also be distorted and non-sinusoidal, despite the sinusoidal shape of the a.c. component of the anode current.

In contrast to this phenomenon, when the core magnetisation is small the transformer operation takes place along the rising linear part of the magnetisation curve; the voltage swings do not reach the saturation region (Fig. 145*b*); and then the non-linear distortion does not take place. When the core is magnetised to the point of saturation, a considerable reduction of the primary winding inductance also takes place. This causes a reduction of amplification on the lower frequencies and results in frequency distortion.

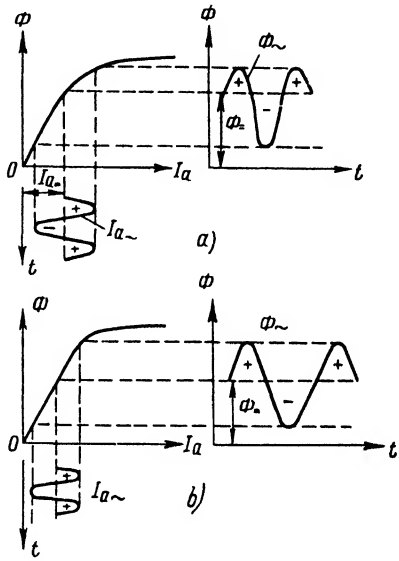


Fig. 145. Non-linear distortion as a result of magnetic saturation

There are several ways of preventing core saturation. One way is that of using valves drawing low anode current. The anode current is reduced by applying bias voltage to the grid of the valve, which, as we already know from Sec. 71, is also necessary for other reasons. Core saturation can be avoided also if the cross-section of the core is made sufficiently large, or else, when an *air gap* is provided in the core. Such an air gap, normally filled with paper, cardboard or some other diamagnetic substance, increases the reluctance of the core, thus reducing the magnetic flux and, accordingly, also reducing the possibility of core magnetisation up to the saturation.

Parallel anode feeding of a transformer-coupled circuit (Fig. 146) offers a radical method of core-saturation-prevention. In this arrange-

ment, known as resistance-transformer coupling circuit, d.c. anode current flows through resistor R_a and, because of dividing capacitor C_d, does not enter the transformer primary winding at all. The capacitance value of the dividing capacitor ranges from 0.1 to 0.5 mfd, which is quite sufficient to let the capacitor pass the a.c. component of the anode current to the primary winding of the transformer. Thus, in such a circuit there is no direct current flow through the transformer primary and the danger of core saturation is completely eliminated.

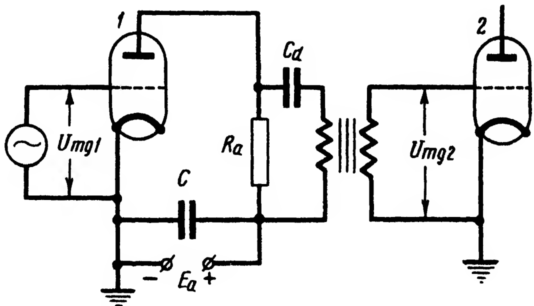


Fig. 146. Transformer-coupled amplifier with a parallel anode feed

The circuit, however, has its own disadvantages. One of them, similar to the shortcoming of resistance-coupled amplifiers, is the loss of a considerable part of voltage, developed by the anode power supply, in the anode resistor R_a.

The other disadvantage of the resistance-transformer coupling is that the amplification of the stage is reduced (in comparison with a standard transformer-coupled stage) because of resistor R_a's connection in parallel with the primary winding of the transformer, the result of which is a decrease of the anode load-resistance.

The resistance-transformer coupling circuit—also called "the rheostat-transformer circuit"—can be so employed that it will artificially secure a lift of the usually attenuated lower frequencies. To do this, the capacitance of the dividing capacitor C_d is so selected that its capacitive reactance x_C is made equal to the inductive reactance of the primary winding of the transformer on some low frequency, for instance — on 50 cps. If this is done, the voltage resonance will occur on the given frequency, boosting up the amplification of this frequency and other frequencies near it, and compensating the fall of amplification on these frequencies which may occur in other stages of the amplifier.

The described method of compensation of attenuated frequencies levels out the frequency characteristic of the amplifier as a whole and is an example of *frequency-distortion-correction*.

In comparison with the described and quite popular resistance-transformer coupling arrangement, other types of combined coupling circuits (resistance-choke and choke-transformer circuits) have found a much smaller application.

Below are discussed some general aspects of the interstage transformer construction.

Interstage transformers, as a rule, employ shell cores, i.e., such cores in which the magnetic flux is split into two branches and the windings are placed on a common coil form. Core laminations, of which

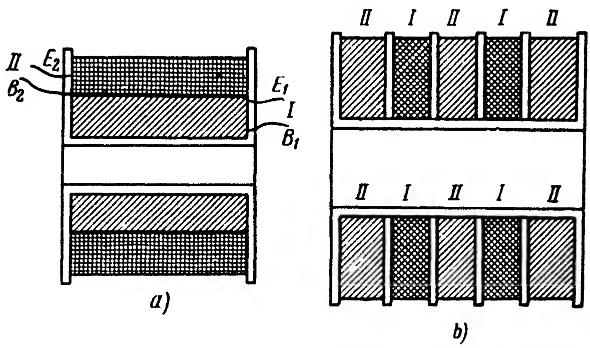


Fig. 147. Cylindrical and sectionalised windings

the transformer core is assembled, resemble in shape Russian letter Ш (shaa) and their sizes, are, accordingly, denoted as Ш-20, Ш-25, Ш-30, etc, where the figure indicates the width of the central leg of the lamination in millimetres. The windings can be of cylindrical, or else, of sectionalised type, the first type (Fig. 147*a*) being more frequently encountered. Either the primary or the secondary winding can be wound next to the core. Sectionalised windings (Fig. 147*b*) are used when it is desired to keep the distributed interturn capacitance and the leakage inductance down to the smallest possible value. In this case, the winding is split up into several sections, the sections of one winding interleaving with the sections of the other winding.

A reduction of the distributed capacitance of a transformer can be secured by the proper connection of the winding terminals, when the terminals closest to each other are connected to $+E_a$ and to the cathode. For instance, if the windings are located as shown in Fig. 147*a*, the terminals which are nearest each other are represented by the end (E_1) of the primary winding and by the beginning (B_2) of the secondary winding. Circuit connections should be so made that the beginning (B_1) of the primary winding is connected to the anode, its end being connected to $+E_a$ of the anode power supply. The beginning (B_2) of the secondary

winding should be connected to the cathode, its and (E_2)—to the grid.

Parasitic inductive and capacitive couplings between transformer windings and the other components of a radio circuit are eliminated by enclosing the transformer in a steel shielding case, which, together with the transformer core it contains, should be earthed to the chassis.

75. GRID BIAS VOLTAGE IN AMPLIFIERS

Negative grid bias voltage is used in amplifiers *for the purpose of shifting the operating point to the left on the characteristic, when the amplifier is to operate without grid current, and also for the purpose of reducing the anode current.*

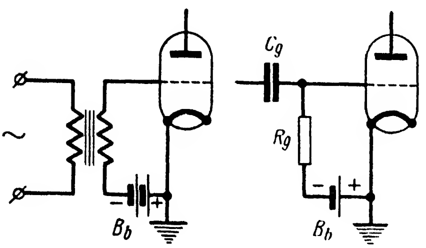


Fig. 148. Grid bias supplied from a separate source

Shifting the operating point precludes a flow of the grid current and, thus, decreases the non-linear distortion of the amplifier. Reduction of the anode current serves the following purposes: it economises the energy of the anode power supply, reduces the anode heating, decreases voltage drop and the associated energy loss in the anode load rasistor R_a in resistance-coupled amplifiers, and reduces the saturation of iron cores in transformer-coupled and choke-coupled amplifiers. The necessary value of bias voltage is determined by the position of the operating point on the valve characteristic. When such position is found, the correct bias voltage E_b and the resting anode current I_{a0} become immediately known (see Sec. 71). Below are discussed various methods of generating the bias voltage and of feeding it to the grid circuit of a valve.

Separate bias supply. The simplest way to obtain the bias voltage is to connect an independent voltage supply between the grid and cathode of a valve, the negative terminal of such supply being connected to the grid (through the grid circuit components), the positive — to the cathode. An ordinary dry-cell battery or else a storage battery can be used as the bias voltage supply. High-power amplifiers can employ batteries, types БАС-60 or БАС-80, or else a special low-power rectifier.

Fig. 148 shows the connection of a separate bias supply to a transformer-coupled amplifier (the bias voltage is fed to the grid through the secondary winding of transformer); it also shows the connection of a similar supply to a resistance-coupled or choke-coupled amplifier (the bias voltage is fed to the grid through grid resistance R_g).

In such circuits, no current is drawn from the bias-supply for the simple reason that the grid of the valve does not consume any current. Under such conditions, even a small dry battery, used as bias-supply, will operate for long periods of time.

The grid bias battery is shunted with sufficiently high capacitance in order to exclude the internal resistance of the battery from the circuit (the capacitor is not shown in Fig. 148). The constancy of grid bias voltage and the independence of this voltage of the operating condition of the valve are the advantages of such bias supply arrangement. This is the reason why separately-supplied bias is sometimes called *fixed bias*. The only disadvantage of such biasing arrangement is the necessity of having a separate bias supply source.

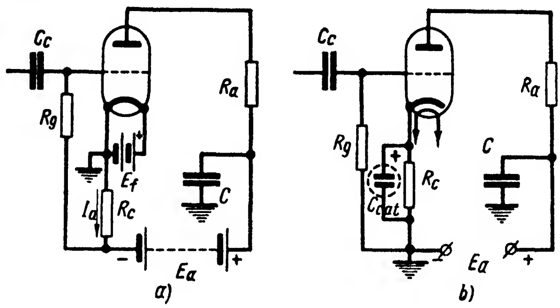


Fig. 149. Automatic grid bias supply used by cathode-type and filamentary valves

Automatic grid bias. Modern amplifiers and radio receivers widely use the so-called automatic biasing arrangement, also known as self-bias. Self-biased circuits obviate the necessity of external sources of bias supply and develop their own bias voltage by virtue of anode current flow through a resistor. This method of obtaining the bias voltage utilises a small part of voltage developed by the anode power supply. Fig. 149 shows how the automatic grid bias is obtained in circuits employing filamentary and cathode-type valves.

Here, resistor R_c, called the *biasing resistor* or the *cathode resistor*, is connected between the negative terminal of anode supply and the filament or cathode of the valve. The grid circuit, i.e., the wire from the secondary winding of the interstage transformer, or the wire from the grid resistor R_g, is connected to the negative terminal of the anode power supply E_a. In such a case, the metal chassis of radio receiver or amplifier is usually connected to $-E_a$ (if the set uses indirectly-heated valves) or to $-E_f$ (in sets with filamentary valves). Thus, bias resistor R_c is connected at the same time to the grid circuit and anode circuit. The d.c. component $I_{a=}$ of the anode current, flowing through resistor R_c, produces a voltage drop across

this resistor.* The end of the resistor connected to $-E_a$ assumes a negative potential in respect to the other end, which is connected to the cathode (or to the filament). Thus, the grid of the valve will become negative in relation to the cathode, the value of this negative voltage being determined by the voltage drop built up across R_c by current $I_{a=}$. This is the bias voltage and its value is determined from the Ohm's law in the following way:

$$E_b = I_{a=} R_c.$$

For instance, if $R_c = 500$ ohms and $I_{a=} = 4$ ma (i.e., 0.004 a), the bias voltage is given by the following:

$$E_b = 0.004 \times 500 = 2 \text{ v.}$$

In most types of radio equipment, the value of R_c ranges from several hundred to several thousand ohms. This resistance is usually of the wire-wound variety, which secures the constancy of its resistance value.

In circuit design practice it often becomes necessary to calculate the value of R_c to give a definite value of bias voltage E_b. Of course, to do this, the value of the d.c. component $I_{a=}$ of the anode current must first be determined. An example of the calculation is as follows.

Suppose that the operating point on the characteristic determines the value of bias voltage E_b as 5 v, and the value of $I_{a=}$ is equal to 2 ma (i.e., 0.002 a). Then, the value of R_c is given by the following equation:

$$R_c = \frac{E_b}{I_{a=}} = \frac{5}{0.002} = 2{,}500 \text{ ohms.}$$

The value of voltage developed by the self-biasing circuit is the part of the anode supply voltage consumed by resistor R_c. In accordance with this, the anode voltage of circuits employing the automatic grid bias is always decreased by a value equal to the bias. If the anode voltage U_a of an ordinary resistance-coupled amplifier with fixed bias supply is smaller than the voltage of the anode supply E_a by the value of the voltage drop across the anode load resistor, i.e., $U_a = E_a - I_{a=} R_a$, the anode voltage in the same amplifier, now provided with the automatic bias, will be still less by the value of voltage drop in R_c, as given by the following:

$$U_a = E_a - I_{a=} R_a - I_{a=} R_c.$$

For instance, if $E_a = 160$ v, $R_a = 40{,}000$ ohms, $R_c = 5{,}000$ ohms and $I_{a=} = 2$ ma (i.e., 0.002 a), the anode voltage in such amplifier-stage will be given by the following:

$$U_a = 160 - 0.002 \times 40{,}000 - 0.002 \times 5{,}000 =$$
$$= 160 - 80 - 10 = 70 \text{ v.}$$

* In the case of tetrodes, pentodes, or more complex valves it is not the anode current I_a, but the cathode current I_c which flows through the resistor R_c.

Usually, E_b is many times smaller than U_a. Therefore, a small decrease of U_a because of using up a part of E_a for the automatic bias is of no great significance.

Operation of the automatic bias circuit usually brings up difficult, problems for the students, although the system is, really, quite simple to understand.

Let us first analyse the automatic bias arrangement from the point of potential distribution at various points of the circuit. Assume a case when the chassis of the amplifier is connected to $-E_f$ (Fig. 149*a*). Let $R_c = 400$ ohms anf $I_{a=} = 5$ ma $= 0.005$ a. The voltage drop across R_c will be equal to $0.005 \times 400 = 2$ v. The chassis, the cathode and the end of R_c to which they are connected (the top end in the circuit diagram) are at zero potential. The other end of R_c, connected to $-E_a$, is at a 2-volt negative potential because the voltage drop across R_c is equal to 2 volts. This -2 v potential is fed to the grid through resistor R_g; consequently, the gris also has a negative potential of 2 volts in respect to the cathode.

Fig. 149*b* illustrates the case when the negative terminal $-E_a$ of the anode power supply is connected to the chassis. Here, the lower end of R_c is the point of zero potential, while the other end of this resistor, connected to the cathode, has a potential which is two volts higher than the chassis, i.e., has a potential of $+2$ v. Thus, the cathode has a potential of $+2$ v, while the grid — connected through R_g to the point of zero potential — is at the zero potential. In the study of various electronic processes taking place in a valve, it is important to know the potential of the grid in respect to the cathode. This is logical, because a bias voltage is nothing but the difference of grid and cathode potentials. In the case here analysed, the bias voltage of the valve is given as

$$E_b = 0 - (+2 \text{ v}) = -2 \text{ v}.$$

Thus, regardless of the point to which the chassis is connected, i.e., regardless of the fact which point of the circuit is considered to be at zero potential, the automatic bias circuit places the grid at a negative potential in respect to the cathode.

The absence of a separate source of bias is the advantage of the automatic bias circuit. Its disadvantage is the inconstancy of bias voltage, which is attributed to the following. The voltage drop across resistor R_c depends upon the value of the d.c. component $I_{a=}$ of the anode current. This value changes when the operating condition of the valve varies, i.e., when the voltages of the filament, anode and screen grid circuits are changed. The change of the $I_{a=}$ component, in turn, changes the grid bias voltage of the valve.

It is interesting to note that such instability of grid bias voltage is useful in certain cases. For instance, if for one reason or another the anode voltage has decreased,, the valve characteristic shifts to

the right. This calls for a decrease of grid bias voltage, if the stage is to continue its normal operation.

The required grid-bias-voltage reduction is automatically provided by the circuit here discussed. The decrease of anode voltage will cause a decrease of the anode current, which will bring about a decrease of voltage drop across resistor R_c, i.e., bring about a decrease of the bias voltage.

Since the anode current contains not only the d.c. component, but also the a.c. component, the latter is bypassed by capacitor C_c (Fig. 149*b*). This capacitor shunts the bias resistor R_c and has a large capacitance value, which means that its reactance is several times smaller than the resistance value of R_c. Special low-voltage electrolytic capacitors, having a capacitance of several dozen microfarads, are used for this purpose. The purpose served by the bypass capacitor C_c is that of keeping down the value of a.c. component across R_c. As far as a circuit of type shown in Fig. 149*b* is concerned, this alternating voltage is considered lost, because it is not applied to the grid of the following valve. Besides, this voltage is fed to its own stage in a phase which is opposite to the phase of the input signal. Let us discuss this last statement in detail.

Assume that in the circuit shown in Fig. 149*b* there is no bypass capacitor C_c and a positive half-wave of alternating voltage reaches the grid of the valve. This will cause an increase of the anode current and a consequent increase of the voltage drop across resistor R_c. As a result, the lower end of resistor R_c will deliver a higher negative potential to the grid of the valve, this negative potential partly neutralising the positive potential developed on the grid by the signal. This "bucking" effect reduces the value of the signal-voltage on the grid and reduces the amplification factor of the stage. The name of this effect is the *negative feedback*. It is this negative feedback that the capacitor, connected across R_c, helps to decrease, thus maintaining the amplification of the stage.

Negative feedback, however, has found useful amplification in certain circuits. Modern amplifiers, using special circuits discussed in Sec. 79 of the present chapter, employ the effect of negative feedback to reduce non-linear and frequency distortion.

Filament-developed bias. When the filament battery has higher voltage than that actually required by valve filament, the excessive voltage of the battery can be utilised as bias, if the filament voltage dropping resistor is connected into the negative wire of the battery, as shown in Fig. 150. In this case, the grid circuit terminates at the negative terminal of the filament battery ($-E_f$). Here is a numerical design-example of a circuit of this type. Assume that the normal filament-voltage of the valve is 4 v. If the filament-battery voltage is 5.2 v, then the 1.2 v voltage drop across resistor R will be the bias voltage, the negative polarity of this voltage being applied to the valve-grid through resistor R_g. This method resembles the auto-

matic biasing method. The difference is that the other method employs a part of the anode voltage to generate the bias, while in the method here discussed a part of the filament voltage is used to provide the required bias. This method, however, is unsuitable when the excess voltage of the filament battery is insufficient to give the correct value of bias. Since it would not be a good practice to increase the filament voltage of the battery only for the purpose of biasing, the method here described has found but a limited application.

It is interesting to note that when no biasing circuit is used and the grid circuit of a filamentary valve is connected to the negative terminal of the filament, a certain amount of negative bias will be applied to the grid. To understand this phenomenon, we will have to consider the value of potentials in different points of the circuit.

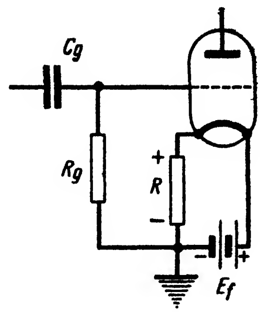


Fig. 150. Filament-developed bias

Assume that a voltage of 2 v is applied to the filament, the negative side of which is connected to the chassis, i.e., serves as the zero-potential point. In such a case, the other end of the filament will have a potential of +2 v, and the middle point of the filament (inside the valve) will be at the potential of +1 v. The grid of the valve is connected to the common negative terminal and its potential is equal to zero. However, its potential is negative in respect to various points of the filament. In respect to the positive end of the filament, the potential of the grid is —2 v, because $0 - (+2) = -2$. In respect to the middle point of the filament, the grid potential is —1 v. And only in respect to the negative end of the filament, the grid is at zero potential. On the average, the grid potential will be —1 v in respect to the filament. Therefore, connecting the grid circuit to the negative end of the filament places a negative bias voltage on the grid, and the value of this voltage will be equal to one-half of the filament voltage.

76. SINGLE-ENDED FINAL STAGE OF AMPLIFICATION

A power amplifier must amplify the power of low-frequency signals with the smallest possible distortion and supply this power to some type of load, usually a loudspeaker. This can be done only when the load has a correct value, which is determined as follows: ***in the case of triodes***, the load resistance R_a must be equal to $2R_i$ (permissible values actually range from $2R_i$ to $3R_i$); ***in the case of pentodes and beam-tetrodes***, the value of R_a may be from $0.05R_i$ to $0.2R_i$. Slight deviations from the stipulated values are, of course,

tolerable. The comparatively small values of R_a used with the pentodes and beam tetrodes are attributed to the characteristics of these valves (see Sec. 72).

If the output power does not exceed a few watts, the final stage (also called the output stage) can employ a single-ended circuit, this name pertaining to an amplifier stage using one valve of the usual triode, tetrode or pentode variety. The final stage may be coupled to the preceding stage (the driving stage, or simply the driver) by means of any coupling arrangement that we have already discussed (transformer coupling, choke coupling, resistance coupling, etc). However, resistance-coupling is used most frequently, because it introduces the smallest distortion. Choke-coupled drivers are seldom employed. The final amplifier is always biased, because the amplitude of signal delivered to its grid can be quite high (several volts and even several dozens of volts). In most cases, this stage uses the automatic biasing circuit.

The circuit connecting the load (for instance a loudspeaker) to the final amplifier is called the output circuit.

There are several types of output circuits, as listed and discussed below.

Direct-output circuit. The loudspeaker can be connected directly into the output circuit, as shown in Fig. 151*a*. Simplicity is the only advantage of this circuit, noted for several serious disadvantages. If the loudspeaker is of low-impedance variety (for instance an electrodynamic speaker), the direct-output circuit simply can not be used because, in this case, the load resistance will be much smaller than the anode resistance of the valve and the stage will give attenuation of the signal applied to its grid, instead of giving amplification. The efficiency of such a circuit will be extremely low. Hence, the direct-output circuit can be used only with high-impedance loudspeakers.

Another disadvantage of the direct-output circuit is the presence of high-voltage direct current at the loudspeaker or at the earphones. When a loudspeaker or a pair of earphones are connected into such a circuit, the polarity must be strictly observed; a wrong connection will cause demagnetisation of speaker or earphone magnets.

Continuous presence of high voltage at the load connected to the direct-output circuit is the most undesirable and at times quite dangerous feature of this circuit. If a pair of earphones are connected into such a circuit as the load, and if the insulation of the earphones is defective, the person using these earphones can receive a painful and even lethal shock, particularly if his body is in contact with some metal part of the radio equipment.

It is absolutely prohibited to resort to direct-output circuit when the amplifier is to feed a wire line loaded with loudspeakers and earphones, as these devices would then be placed at a high-voltage potential in respect to earth, thus constituting a general hazard

to the users. Apart from this consideration, which, quite naturally, comes first, direct coupling of an amplifier to a sound-diffusion line is undesirable also because of the following technical considerations: Sound-diffusion lines are usually noted for large values of leakage current to earth. If such lines are made to carry high voltage, the leakage would become excessive, resulting in a large waste of the electric power. Another objection is as follows. When a great number of loudspeakers—even high-impedance loudspeakers—are

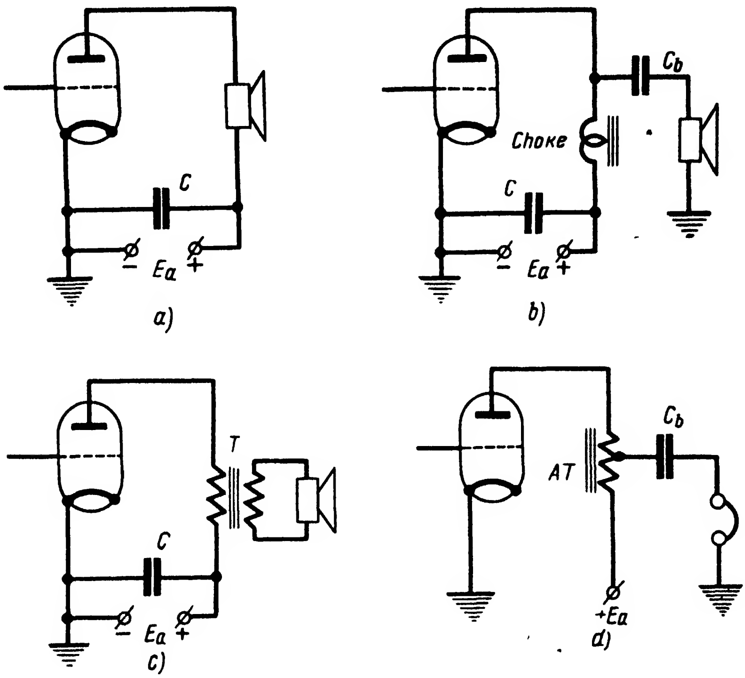


Fig. 151. Amplifier output versions: *a)* direct-coupled; *b)* choke-coupled; *c)* transformer-coupled; *d)* autotransformer-coupled

connected to a sound diffusion line in parallel, the impedance of such line becomes very low. If the line is connected directly into the high-impedance anode circuit of an amplifier, the amplifier will not give any amplification.

Because of all the above-named disadvantages, the direct-coupled circuit has found only very limited application, for instance, in the coupling of high-impedance earphones and loudspeakers to very simple radio receivers.

Choke-coupled output circuit. In this circuit, the d.c. component of the anode current flows through a low-frequency choke and does not reach the loudspeaker (Fig. 151*b*). The loudspeaker is connected

to the anode of the final-stage valve through a blocking capacitor, which blocks the d.c. component but has sufficient capacitance to pass the a.c. component to the load. Loudspeakers and earphones connected to the choke-coupled circuit are not placed at a hazardous d.c. potential in respect to earth and their electromagnets cannot be demagnetised by a wrong connection to such a circuit. However, this type of circuit still cannot be used for feeding low-impedance loudspeakers or sound diffusion lines, because the low-impedance load connected through blocking capacitor C_b across the choke would prevent the amplifier from normal functioning. This is why choke-coupled output circuit is only seldom used and then, chiefly, in such application as feeding auxiliary loudspeakers (Fig. 152).

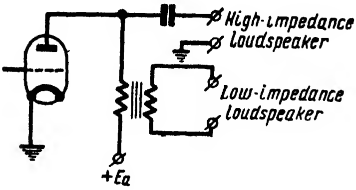


Fig. 152. Connection of a final stage to loudspeakers with different values of impedance

Transformer-coupled output circuit. All the disadvantages listed above and pertaining to direct-output circuit and to choke-coupled output circuits are eliminated in the transformer-coupled output circuit, pictured in Fig. 151*c*, which is nothing but the transformer-coupled amplifier circuit previously studied in Sec. 74. Transformer-coupled output circuit, or more simply—transformer-output, is the most popular and widely used of all output circuits employed by final stages of amplifiers.

In this output circuit, the d.c. component of the anode current flows through the primary winding of the output transformer and does not reach the loudspeaker because of the insulation between the primary and secondary windings. The loudspeaker can have any impedance, because the transformer converts the load value. With proper transformation ratio, low-impedance loudspeakers are properly matched to any load.

This property is the cardinal advantage of transformer-coupled output circuit.

Let us now discuss the transformation feature of the given circuit.

Assume that the final stage employs a triode, whose R_i is equal to 1,000 ohms. Then the value of R_a must be 2,000 ohms, because $R_a = 2R_i$ is the most advantageous value of the anode load of a triode. If the value of the actual load (loudspeaker), connected to the secondary winding of the output transformer, is equal to 20 ohms, we take an output transformer with transformation ratio of $n = 10:1$ and obtain the value of resistance which, when related to the primary winding, is expressed as follows:

$$R_1 = R_l n^2 = 20 \times 10^2 = 20 \times 100 = 2{,}000 \quad \text{ohms.}$$

It is customary to say that the output transformer matches the load resistance to the anode resistance of the output valve in accordance with the following equation:

$$R_a = R_l n^2.$$

This formula is transposed and used as follows when it is desirable to determine the transformation ratio from known values of R_a and R_l:

$$n = \sqrt{\frac{R_a}{R_l}}.$$

For instance, if $R_a = 6{,}000$ ohms and $R_l = 15$ ohms, then:

$$n = \sqrt{\frac{6{,}000}{15}} = \sqrt{400} = 20.$$

Low-impedance loads require the application of step-down transformers in the output-stages of amplifiers. Depending upon the type of valve used and upon the resistance of the load, the transformation ratio can be of different value. The primary winding of the output transformer usually contains from 2,000 to 6,000 turns, so that its inductive reactance on low frequencies would not be too small to cause considerable decrease of amplification.

The cross-sectional area of the output transformer core depends upon the power level handled by the stage. Cores with cross-sectional areas of 1 sq cm and larger are commonly used. An air gap is sometimes provided in the core in order to reduce the magnetisation of iron by the direct current flowing through the primary winding of the transformer to the anode of the valve. The windings of the output transformer must have the smallest practicable value of distributed capacitance. The diameter of the wire used in the windings must be correctly calculated in accordance with the current density and will be normally larger than the diameter of wire used in interstage transformers.

When an output transformer is to feed low-impedance and high-impedance loudspeakers, it is sometimes provided with two secondary windings. One of these windings consists of a small number of turns and is designed for connection to low-impedance loads, while the other winding, designed to feed high-impedance loads, is comprised of a large number of turns. An alternative and a simpler way of solving the same problem is shown in Fig. 152. In this circuit, a simple output transformer with a single secondary winding is employed, the latter being designed to feed low-impedance loudspeakers. High-impedance loudspeakers are connected to the stage by means of a choke-coupled circuit in which the function of the choke is performed by the primary winding of the transformer. The secondary winding of an output transformer is sometimes provided with taps to permit the selection of an optimum transformation

ratio for various types of load. Tapped output with an appropriate tap-selecting switch is required by all amplifiers working into sound diffusion lines, because the resistance of load connected to such a line depends upon the number of subscriber-loudspeakers used at the moment.

Autotransformer-coupled output circuit. Final stages sometimes employ autotransformers instead of output transformers (Fig. 151*d*). Output autotransformers have all the advantages of the output transformers but they are not convenient on low-impedance loads, where the blocking capacitor must have a very low capacitive reactance (a few ohms), which means that its capacitance has to be

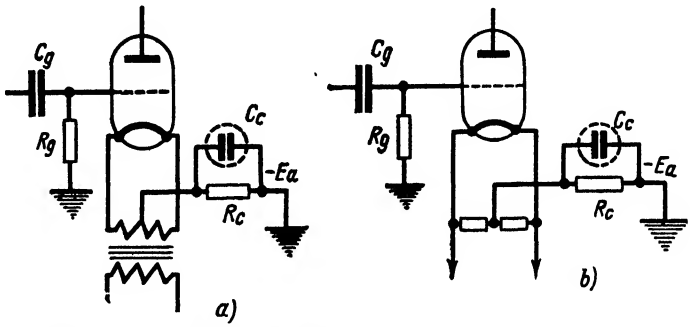


Fig. 153. Zero-point arrangements in filament circuits

hundreds of microfarads. Autotransformer-coupled output circuits are sometimes used by radio receivers to drive high-impedance loads whose impedance has to be matched to the output impedance of the final stage. Compared to an ordinary output transformer of similar power rating, an autotransformer has smaller dimensions, because its core uses less iron, and the winding — less copper.

Let us now turn our discussion to various valves used in the output stages of amplifiers.

Special high-power triodes have amplification factor of about 4-10, anode resistance of 500-2,000 ohms and "left-handed" characteristics. Some of these triodes are of the filamentary type, but their filaments are thick, have a considerable thermal inertia and may be fed with alternating current. A.c. hum, i.e., pulsations of the anode current at the frequency of 50 cps, heard in a loudspeaker as a peculiar humming sound, can be reduced by providing a so-called *zero point*, to which are connected both the anode and the filament circuits. Such zero point may be represented by the centre tap of the filament winding of the power transformer or else by the centre tap of a divider (potentiometer) having a resistance of a few dozen ohms and connected in parallel with the filament circuit (Fig. 153).

Besides various types of triodes, pentodes and beam tetrodes are also widely used in the output stages of amplifiers. The pentodes and tetrodes have higher values of amplification factor and anode resistance than triodes, these values reaching $\mu = 50\text{-}200$ and $R_i = 20{,}000\text{-}200{,}000$ ohms. Because of the high value of μ, the grid circuits of pentodes and tetrodes require a much smaller signal voltage to drive them to full output. Hence, an amplifier using these types of valves will have a smaller number of stages than a similarly rated amplifier employing triodes. However, pentode and tetrode amplifiers introduce somewhat greater non-linear distortion than triode amplifiers.

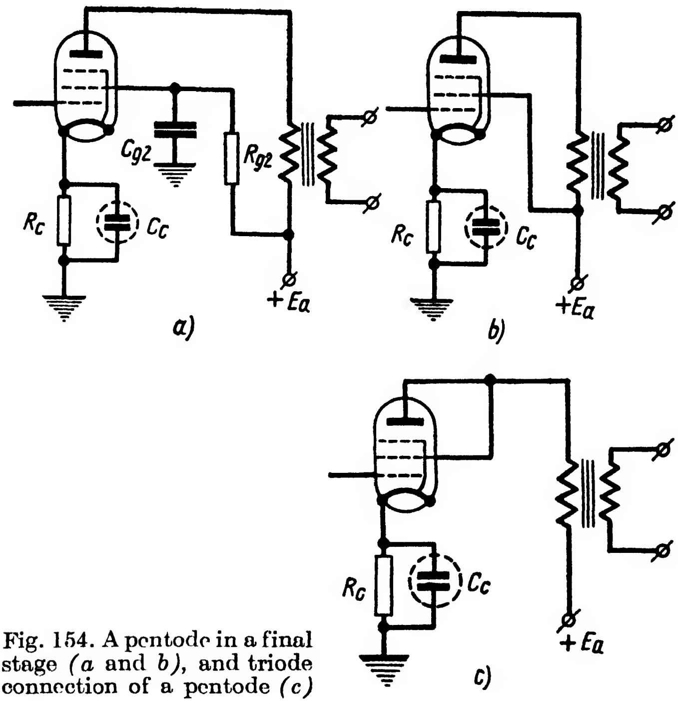


Fig. 154. A pentode in a final stage (*a* and *b*), and triode connection of a pentode (*c*)

Output transformers used with pentodes and tetrodes have a greater number of primary winding turns and a higher transformation ratio, in comparison with output transformers employed by triodes. This is explained as follows. Triodes have anode resistance values from 500 to 2,000 ohms and the resistance of the load into which they work must be between 1,000 and 4,000 ohms, if the optimum operating condition $R_a = 2R_i$ is to be observed. For pentodes and beam tetrodes, in which $R_i = 20{,}000\text{-}200{,}000$ ohms, the anode load value is from 1,000 to 40,000 ohms, basing on $R_a = 0.05R_i - 0.2R_i$.

In order to secure "left-handed" characteristics of pentodes and beam tetrodes, the screen grid voltage must be equal to 70-80% of the anode voltage (Fig. 154*a*) and in some cases must be even of the same value as the anode voltage (Fig. 154*b*). The dropping resistor in the circuit of the screen grid R_{g2} must be designed to carry a considerable current. Capacitor C_{g2} of the same circuit must have a value of at least 2 mfd, so that its capacitive reactance on the lower frequencies is much lower than the resistance value of R_{g2}.

In some output stages pentodes and beam tetrodes are used as triodes (Fig. 154*c*). This is resorted to when the maximum amplification that a valve can give is not required, and when it is desirable to reduce non-linear distortion as much as possible.

Triodes connection of a pentode valve should not be confused with the case when the screen grid is supplied with the same value of voltage as the anode (Fig. 154*b*).

When a higher level of output power is required, two or three valves are connected together and used in the final stage (this is anode-to-anode and grid-to-grid connection). In such a circuit, voltage values remain the same as in the case of a single-valve circuit, but the current and power levels are correspondingly multiplied (doubled in case of two parallel-connected output valves). It may be considered that the valves thus connected in parallel are equivalent to a single valve in which μ remains the same, mutual conductance S is increased, while the anode resistance R_i is decreased by the number of times equal to the number of parallel-connected valves. If two valves are employed in such a circuit, the equivalent parameters will be, correspondingly, μ, $2S$ and $0.5R_i$. Actually, in a circuit of this type, because of slight dissimilarity of the valves, the power increase is not strictly proportional to the number of connected valves, but is somewhat less. Besides, when the valves are connected in parallel, the interelectrode capacitances are accordingly increased, because they are connected in parallel. This is undesirable, because the increase of the input capacitance of the final stage results in a more pronounced attenuation of the higher audio frequencies in the driver stage. Because of this, it is not advantageous to connect in parallel more than 2-3 valves. When a higher power output is desired, it is a better practice to employ a single valve of higher rating, or else to use a push-pull stage.

The dynamic characteristic of a valve used in the final stage of an amplifier and employing transformer output has certain peculiarities. The transformer presents high impedance only to the a.c. component of the anode current. On the other hand, the d.c. resistance of the primary winding of transformer is very small and can be disregarded. Hence, in transformer-coupled amplifiers the operating point corresponds to the anode voltage under resting condition, i.e., corresponds to the voltage E_a of the anode power supply, and not to a smaller voltage value (as it does in a resistance-coupled amplifier).

When the anode dynamic characteristic is plotted for a transformer-coupled amplifier, the operating point A is plotted first (Fig. 155). The position of this

point is determined by the selected grid bias value E_b and by the anode supply voltage E_a. Resting current I_{a0} corresponds to this point. The second point of the dynamic characteristic can be determined from the equation $U_a = E_a -$ $- \Delta I_a R_a$. In this equation R_a is the load resistance related to the primary winding, i.e., $R_a = R_L n^2$, while ΔI_a stands for the anode current change. Here, in contrast to a resistance-coupled amplifier (see Sec. 36), the voltage drop across R_a is produced only by the a.c. component of the anode current and, because of this, the equation does not include the anode current but rather its change ΔI_a.

Assuming $U_a = 0$, we find from the given equation that $\Delta I_a = \frac{E_a}{R_a}$. This value should be added to the current I_{a0} and then we obtain point M on the vertical axis. A line drawn through this point and through point A is the dynamic characteristic.

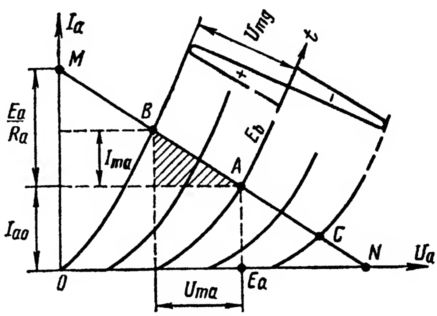


Fig. 155. Dynamic anode characteristic of a transformer-coupled amplifier

Thus, in contrast to the case of a resistance-coupled amplifier, the plotting of the anode dynamic characteristic for a transformer-coupled amplifier implies laying off the value of $\frac{E_a}{R_a}$ from current value I_{a0}, rather than from 0. For smaller or greater values of R_a, the dynamic characteristic assumes a smaller or greater slope, but passes through point A. Point M does not correspond to any real operating condition, because when $U_a = 0$ the anode current cannot have a larger value. Point N corresponds to a real operating condition—to cutting off the anode current by a high negative grid potential.

An interesting peculiarity of a transformer-coupled amplifier is seen in the following. When the anode current is decreased below the I_{a0} value, the anode voltage becomes higher than the value of E_a (AN section). This phenomenon, although incredible at first sight, is explained as follows. The primary winding of the transformer possesses a certain value of inductance and, consequently, when the current decreases or increases in this winding, an e.m.f. of self-induction is built up in it. The polarity of this induced e.m.f. will be opposite to the polarity of the change in current. When the current is increased, this e.m.f. bucks it and is subtracted from voltage E_a, thereby causing the anode voltage to decrease. On the other hand, when the current is decreased, the e.m.f. of self-induction acts in the same direction with the current and is, therefore, added to voltage E_a of the supply source. As a result of this, the anode voltage is increased.

If we are to obtain amplification with tolerable value of non-linear distortion, we must take into consideration, in the case just described, all the instructions given above during the discussions on the dynamic characteristic for resistance-coupled amplifiers, the selection of the operating point, and the selection of the dynamic characteristic slope (i.e., pertaining to the selection of the R_a value) which took place on the subject of resistance-amplifiers. When following the recommended procedure, do not try to obtain the maximum possible amplified voltage, but rather strive to secure the maximum useful power with little non-linear distortion. The value of this power is equal to one half the product of the amplitude values of a.c. components of anode voltage and anode current, i.e.,

$$P = 1/2 I_{ma} U_{ma}.$$

For the amplitude of the alternating grid voltage shown in Fig. 155, the useful power is represented by the area of the shaded triangle (provided that the non-

linear distortion is small, i.e., $AB = AC$). Thus, if we wish to obtain a higher level of the useful power, strive to secure the maximum area of this triangle. In case of pentodes and beam tetrodes, the value of $R_{a\ opt}$, corresponding to the minimum non-linear distortion, also corresponds, as a rule, to the maximum useful power.

The plotting of the dynamic characteristic, discussed above, pertains, as well, to the circuits with choke-coupled and autotransformer-coupled output. It also pertains to high-frequency amplifiers working into a tuned-circuit load. In these cases, the d.c. resistance of the anode load is also very low and, hence, it is permissible to consider $U_a \approx E_a$ under the resting condition.

77. DOUBLE-ENDED OR THE PUSH-PULL OUTPUT STAGE

Push-pull circuit — sometimes called the "symmetrical circuit" — discussed below, gives a good account of itself when employed as the final stage of an amplifier. In comparison with a single-ended amplifier, a push-pull amplifier can develop a much higher power output, accompanied by smaller distortion. With the increase of signal voltage at the grid of a single-ended amplifier, the distortion at the output of the amplifier sharply increases for the following reasons. Grid voltage swings begin to reach the lower bend of the characteristic, and also to enter the grid current region. Magnetic flux changes begin to enter the region of magnetic saturation, as a result of the output transformer-core magnetisation by the large d.c. component of the anode current. These causes of non-linear distortion were discussed in Secs 71 and 74.

In a push-pull circuit the non-linear distortion is decreased because of a different operating principle of the circuit. Fig. 156*a* shows a push-pull output stage employing two triodes. The circuit is, in effect, a twin amplifier stage, consisting of two single-ended stages with common supply sources.

The secondary winding of the input-transformer (also called the grid transformer) is provided with a centre-tap. The primary winding of the output (or the anode) transformer is similarly centre-tapped. The two valves operate with a 180° phase-shift. If an a.c. signal is applied to the primary winding of the grid transformer, the polarity of potentials appearing at points *1* and *2* of the secondary windings will be opposite at any moment in respect to the polarity of the centre-tap, the latter being connected to the cathodes of the two valves through resistor R_c. For instance, if at a certain moment point *1* and the grid of valve V_1 are positive in respect to the cathode, point *2* and the grid of valve V_2 are negative in respect to the cathode at this moment. Alternating voltages on valve grids are shown graphically in Fig. 156 *b* and *c*, while the graphic representation of the anode currents of these valves is given in Fig. 156 *d* and *e*.

The d.c. components of the two anode currents flow through the halves of the output transformer primary winding in opposite

directions, as indicated by arrows in Fig. 156*a*. Because of this, the magnetising action of the d.c. component of anode current of one valve is neutralised by the current of the other valve. As a result, the output transformer core is completely free from constant magnetisation. The danger of valve characteristic entering the sector of magnetic saturation is, thus, fully eliminated and the non-linear distortion is reduced.

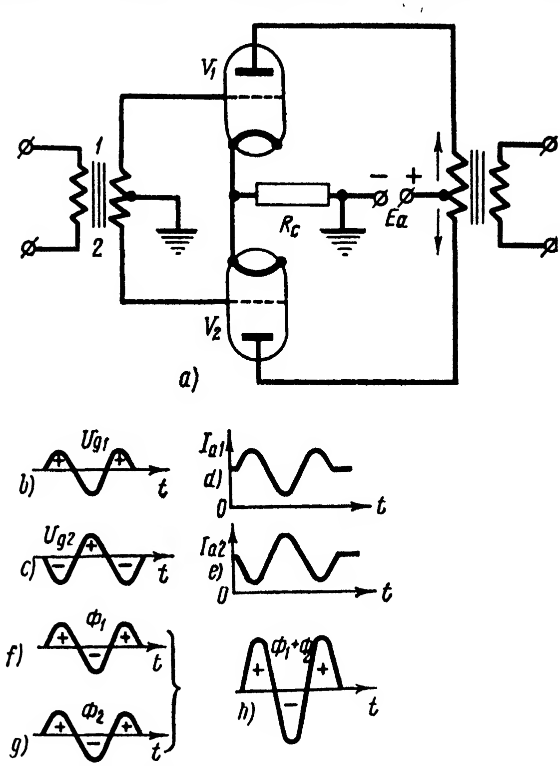


Fig. 156. Circuit diagram of a push-pull final amplifier stage and the curves illustrating its performance

The absence of constant magnetisation makes it possible to considerably reduce the output-transformer core dimensions. Because of a certain asymmetry of the circuit, caused, for instance, by such factors as unequal anode currents of the two valves or slightly different number of turns in the two halves of transformer winding, the core of a practical transformer of this type is subjected to some constant magnetisation. However, such magnetisation is small and causes no detrimental effects. Still, the asymmetry of a push-pull circuit should be kept down to the smallest possible value.

The a.c. components of the anode currents of the two valves also flow in the opposite directions through the two halves (the "legs") of the primary winding of the output transformer, but since they are shifted in phase by 180°, their magnetising effects are added up. The resultant alternating magnetic flux Φ is equal to the sum of alternating magnetic fluxes Φ_1 and Φ_2 of both legs, i.e., it has twice the value of alternating magnetic flux that would be obtained if a single valve were employed (Fig. 156 *f*, *g* and *h*). Thus, the magnetic flux in the transformer core has the frequency of the signal applied to the input circuit of the stage and induces an alter-

nating voltage of the same frequency in the secondary winding of the output transformer.

The action of the two valves is thus added up in the core of the output transformer, giving a doubled power at the output circuit. Of course, in practical circuits, owing to a certain amount of the inevitable asymmetry, the power output will be slightly less than the doubled value.

In the common part of the anode circuit of the two valves, from the centre-tap of the primary winding of the output transformer to the cathode, the d.c. components of the anode currents of both valves flow in one direction and add up, i.e., the values, viewed as consumers of d.c. energy supplied by the anode power source, are connected in parallel. On the other hand, the a.c. components of the anode currents are mutually cancelled in this part, because they are in phase opposition. The absence of a.c. component in the common part of the anode supply is an important advantage of the push-pull system. Due to this feature of the system, only d.c. voltage exists across the biasing resistor R_c. Hence, there is no negative feedback to reduce the amplification of the stage and there is no need to shunt R_c with a large capacitor. Besides this, the absence of the a.c. component in the common part of the anode circuit eliminates the deleterious influence which the final stage could otherwise impose upon the preceding amplification stages through the power supply (this deleterious influence will be discussed in detail in the following section).

Another advantage of the push-pull circuit is its low sensitivity to pulsations of the power supply voltage. Alternating current supply of the filament circuit and insufficient filtering of the anode supply produce less hum in a loudspeaker connected to a push-pull amplifier, in comparison with the hum level produced in a loudspeaker connected to a single-ended amplifier. This is attributed to the following. In the push-pull circuit the valves are connected in parallel as far as the feed circuits are concerned, and the feeding currents pulsate in the same phase under the influence of feed-voltage pulsations. But the a.c. components imposed upon the anode currents by these pulsations (and also the d.c. components) flow in opposite directions through the two legs of the output transformer winding and their magnetic fields mutually cancel each other. Only a small ripple (pulsation) remains because of the asymmetry of the circuit.

A push-pull circuit can work under special operating conditions, which are quite intolerable in a single-ended circuit, but which provide higher power output and higher efficiency.

In accordance with the condition of their operation, push-pull amplifiers are grouped into several classes, discussed below.

Class A amplifier is an amplifier which operates on the linear part of its characteristic, where anode current oscillations almost exactly

correspond to the oscillations of an alternating voltage applied to the grid of the valve.

In our previous studies we have been dealing with just such a type of amplifiers. The preliminary amplifiers and the single-ended output stages, which we have already discussed, always operate as class A amplifiers. A class A amplifier is noted for the following:

1) small non-linear distortion;

2) comparatively low useful-power because only the linear portion of the characteristic is used, this portion constituting but a small part of the whole curve;

3) the d.c. component I_{a-} of the anode current is equal to the resting current I_{a0} and has a considerably higher value than the amplitude of the a.c. component.

The last one of the three listed peculiarities indicates that the input power (the d.c. power fed to the anode circuit of the valve by the anode power supply) is large in a class A amplifier, and that the greater part of this power is lost as anode dissipation. Such anode dissipation is particularly useless when no signal is applied to the grid of the valve and when the whole of the anode power is used for heating the anode. Theoretically, the efficiency of a class A amplifier cannot exceed 40-45%, while in practical circuits the efficiency is even less than this value. Thus, the operating condition of a class A amplifier, which is unfavourable both from the point of useful power output and from the point of efficiency, has brought about the development of another type of amplifier—class B amplifier, explained below.

Class B amplifier is an amplifier in which the operating point is selected in the beginning of the lower bend of the valve characteristic; in such a circuit anode current pulses occur only on the positive half-waves of alternating voltage applied to the grid of the valve.

Such operating condition is set by an appropriate increase of grid-bias voltage and by a corresponding increase of excitation, i.e., of the alternating voltage applied to the grid of the valve. The graphic representation of operation of one of the valves of a class B amplifier is shown in Fig. 157*a*. In contrast to the shape of grid voltage oscillations, the oscillations of the anode current are so distorted that — were such an operating condition applied to a single-ended amplifier — the harmonic coefficient would reach 40% and even more.

But a class B push-pull amplifier is capable of giving normal performance under such unusual operating conditions. Figs. 157*b* and 157*c* give a graphic representation of anode currents of two valves employed in such an amplifier. The valves work alternately — first one, then the other. The pulsating anode currents of the valves flow in the opposite directions in the two legs of the output transformer primary winding. These currents set up magnetic fluxes in the transformer core, such fluxes changing in a similar manner to the

currents (i.e., in accordance with the curves of Figs. 157*b* and 157*c*) but flowing in opposite directions. Because of this, the resultant magnetic flux will be only slightly distorted (Fig. 157*d*). The shape of voltage, induced by this flux in the secondary winding, will also be only slightly distorted. Hence, although intolerable distortion appears in each leg of the circuit, yet, thanks to the double-ended system, the distortions of one leg are compensated by the other and the output voltage as a whole is only slightly distorted. Owing to

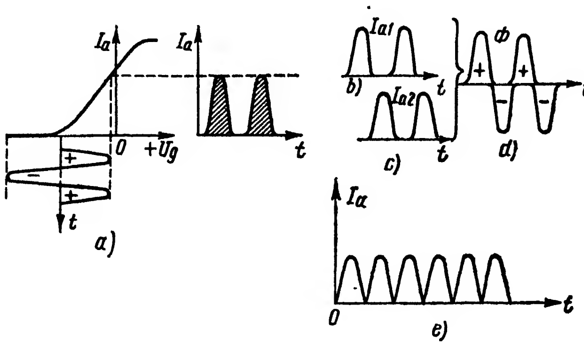


Fig. 157. Operating condition of class B push-pull stage

a certain asymmetry of practical circuits and to the curvature of valve characteristics, the harmonic coefficient can reach 10-15% at the output of an average class *B* amplifier.

Fig. 157*e* shows the anode current in the common sector of a class *B* amplifier. The pulsation frequency of this current is doubled, i.e., the current does not contain the a.c. component with the frequency of the voltage fed to the input circuit of the stage (assuming that the circuit is perfectly symmetrical). This gives the advantages which have already been discussed. When dealing with class *A* push-pull amplifiers it is still, however, necessary to shunt the self-biasing resistor with a high-capacitance capacitor to prevent the building up of an interfering doubled-frequency voltage across this resistor.

Thus, amplification performed by a class *B* amplifier gives higher useful power, compared with class *A* amplification, because the amplifier utilises a greater portion of the valve characteristic. It is true, that the operation of a class *B* amplifier produces greater non-linear distortion than that of a class *A* amplifier, but such distortion still remains within tolerable limits. An important advantage of a class *B* amplifier is its comparatively small consumption of anode supply energy. In this type of amplifier the resting current I_{a0} is very small and in some cases is even equal to zero. This means

that when no signal is applied to the grid circuit of class B amplifier, the latter practically consumes no anode current.

The d.c. component I_{a-} is also comparatively small in such amplifier; it is considerably smaller than the amplitude of the a.c. component. Hence, when a signal is applied to the grid circuit of a class B amplifier, its anode-supply energy consumption also remains lower than that of a class A amplifier. Power dissipated by anodes of a class B amplifier is also smaller at all times, becoming nearly zero during pauses. As a result, the efficiency of class B amplifiers is much higher than that of class A amplifiers and can reach 60-75%.

It should be noted that class B amplifiers require fixed bias and cannot employ automatic biasing arrangement. This is easily understood if we consider that the resting current of class B amplifiers is extremely small (sometimes zero) and is not capable of building up sufficient voltage drop across the biasing resistor. Moreover, the constantly changing value of the anode current d.c. component in a class B amplifier would produce a constantly varying value of bias voltage, if the self-biasing arrangement were employed in such an amplifier.

Besides class A and class B amplifiers, an intermediate class, known as class AB, is also employed. In a class AB amplifier, the operating point is set in the lower bend region, but the resting current is not equal to zero. Signals of considerable amplitude are fed to the input circuit of such an amplifier in order to utilise a greater portion of valve characteristic. In class AB amplifiers, negative half-waves of the signal applied to the grid circuit cause a negative half-wave of anode current; these anode current half-waves are, however, much smaller than positive anode current half-waves. Each leg of a class AB amplifier creates strong non-linear distortion, but, owing to the push-pull arrangement of the circuit, such distortion is cancelled at the output. Single-ended amplifiers cannot operate under the conditions of a class AB stage.

There are two varieties of class AB amplifiers, known, respectively, as class AB_1 and class AB_2 amplifiers. The operating conditions for one of the two push-pull valves used by such amplifiers are given in Fig. 158 *a* and *b*. The operating condition of a class AB_1 amplifier is noted for the absence of grid current, i.e., the stage operates in the negative grid voltage region. The operating condition of a class AB_2 amplifier (Fig. 158*b*) is so set that the valve grid voltage swings into the positive region, thus making the grids draw current. A class AB_2 amplifier gives a higher power-output than a class AB_1 amplifier because it utilises a greater portion of valve characteristic, but it also gives higher distortion.

When indicating the operating condition of a valve used in an amplifier stage, it is customary to attach, respectively, index 1 or index 2 to show that the valve operates without or with grid current. Thus, operating conditions A_1 and A_2, as well as B_1 and B_2 are possible.

Push-pull output stages employ power triodes, pentodes, and beam tetrodes. In such stages, screen-grid voltage is supplied through a common or individual dropping resistors, or else is fed to the screen grids without dropping, directly from the anode voltage source. When high amplification is not sought, pentodes and beam tetrodes are connected as triodes. When more output power is desired, sometimes two or three valves are connected in each leg of a push-pull stage. Special double-triodes are manufactured for application in push-pull stages. Some of such valves are made with right-hand characteristics, being designed for operation in class B_2 amplifiers (amplifiers in which grid current flows). Owing to such characteristics, class B operating condition can be set with a low value of grid bias voltage. Fig. 159 gives the circuit diagram of a push-pull stage employing a double-triode.

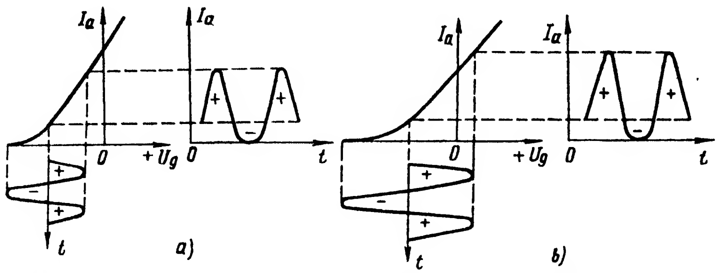


Fig. 158. Operating conditions of class AB_1 and class AB_2 amplifier stages

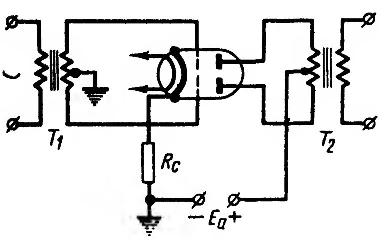


Fig. 159. A push-pull stage employing a double triode

Output transformers of push-pull stages must be symmetric. This is secured by splitting the coil into two sections, over which are wound the halves of the winding provided with a centre tap. The winding is sectionalised to reduce the distributed capacitance and to provide a better protection against electrical breakdown. Such protection is very important in high-power output transformers where the primary winding is at a high potential.

In order to keep down the distortion introduced by transformers, the input transformer is sometimes omitted from push-pull circuits and is substituted by the so-called phase-inverting circuit, also known as phase-shifting circuit. The phase-inverting circuit, or simply phase-invertor, is a resistance-coupled circuit employing one or two valves and developing at its output two equal voltages, opposite in phase and applied to the grids of the final push-pull stage.

Fig. 160 gives two examples of phase-invertors. In the circuit shown in Fig. 160*a* the anode load resistance is divided into two halves (R_{a1} and R_{a2}), connected to the cathode and anode. Amplified alternating .voltages developed across these resistors are fed to the grids of the driven push-pull stage through blocking capacitors C_{g1} and C_{g2}. These voltages are opposite in phase in respect to the point of zero potential (earth). This is a comparatively simple circuit but it functions properly only when the source of the alternating voltage fed to the grid of the driver is not connected to the common negative of the circuit. Such

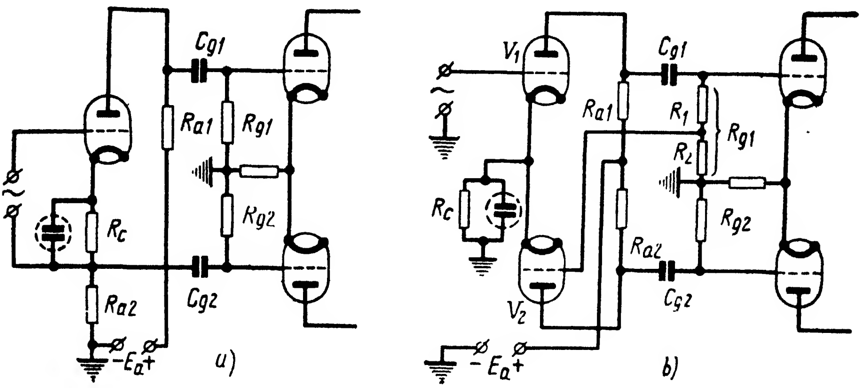


Fig. 160. Phase-invertors

a source may be represented by a pickup or by the secondary winding of a transformer. However, when such a driver is connected to a resistance-coupled preliminary stage, its amplification factor is greatly reduced and can become smaller than two. Such a connection also upsets the symmetry of the circuit.

A better, although more complicated, phase-invertor is shown in Fig. 160*b*; it employs an additional valve. This type of phase-invertor functions by virtue of the 180° phase shift which occurs in any amplifier stage.

Alternating voltage from some source (for instance from a preceding stage) is applied to the grid of valve V_1. Amplified and phase-inverted voltage is developed across resistor R_{a1} and is fed through capacitor C_{g1} to grid resistor R_{g1}, the latter actually comprising two separate resistors R_1 and R_2. Valve V_2, together with its associated components R_{a2}, C_{g2} and R_{g2}, functions as the second auxiliary stage. A part of voltage amplified by valve V_1 is fed from resistor R_2 to the grid of valve V_2. This voltage is amplified by valve V_2, is phase-inverted by 180°, and is applied to resistor R_{g2}. Voltage divider R_1R_2 is so designed, that equal values of voltage appear across R_{g1} and R_{g2}. If, for instance, the amplification factor of the stage using valve V_2 is equal to 10, the value R_2 must be equal to $0.1R_{g1}$. Valves V_1 and V_2 may be of pentode type. In such a circuit it is also convenient to use a double triode. Phase-invertors are used only in such cases where the driven push-pull stage operates as a class A_1 or AB_1 amplifier and draws no grid current.

When the output stage operates with grid current, for instance as a class AB_2 amplifier, power has to be expended to generate the grid current. In such a circuit, the driver stage is called upon to supply a considerably greater power in order to properly excite (to drive) the final stage.

Gain Control

A gain (or volume) control is usually provided at the input of an amplifier. Such a gain control is a potentiometer, across which the a.c. signal voltage is applied. Moving the potentiometer slider, connected to the grid of an amplifier valve, offers a method of varying the amplitude of signal voltage applied to the grid circuit of the given valve.

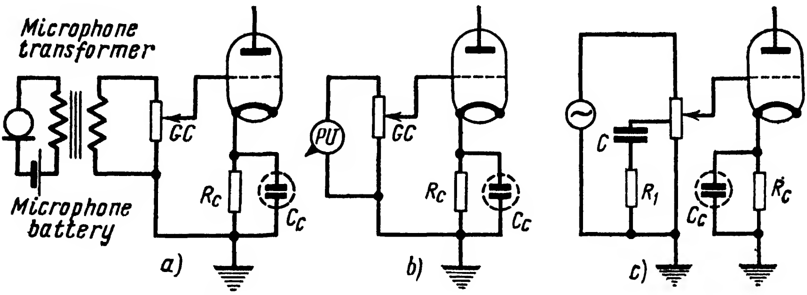


Fig. 161. Tone control connections

Gain control connection circuits, given in Fig. 161 *a* and *b*, pertain to cases when the amplifier operates from a microphone or a pickup. The potentiometer resistance should be several thousands or even several hundreds of thousands ohms. The gain is nearly always controlled at the input of amplifiers in order to avoid overloading the following stages with strong signals, which would give rise to non-linear distortion. It is a poor practice to connect the gain control at a higher level of the circuit, for instance in the output stage, and to try decreasing the amplitude of signals strongly amplified in preceding stages.

When a gain control is adjusted to reduce the signal strength, a peculiar effect becomes noticeable; the lower the gain of the amplifier, the less noticeable become the lower audio frequencies. As a result, the reproduction is distorted at low-signal levels. This effect is attributed to the peculiarity of the ear, which is more sensitive to medium- and higher-frequency sounds than to the sounds of lower frequencies. A special *compensated volume control* (Fig. 161*c*) is used to reduce such apparent distortion. In this type of gain control, a part of the potentiometer is shunted by capacitor C connected in series with resistor R_1.

The capacitor decreases the amplification of the higher-frequency signals and, thus, levels out the audibility on various frequencies. Such compensation is necessary on the reception of music, although it can impair the intelligibility of speech. Capacitor C is, therefore, sometimes short-circuited by a special switch, marked "Music-Speech".

Automatic Biasing Circuits

Amplifiers employing indirectly heated cathode-type valves generally use an individual self-biasing circuit in each stage. Bias resistor R_b is connected into the cathode of each valve, the anode current of the valve developing a required voltage drop (bias) across such resistor (Fig. 162*a*). The resistor is shunted by a capacitor, the latter passing the a.c. component of the anode current. Such automatic biasing variant offers the maximum convenience, because each valve is biased individually and independently of the other valves of the amplifier; in such an arrangement, any required value of bias voltage may be applied to the grid of any valve without affecting the bias voltage of other valves. When an amplifier employs filamentary valves, it becomes impossible to use such automatic biasing arrangement in every stage, because the filaments of all the valves are connected in parallel and are fed from a common power supply. Because of this, such amplifiers have to use a common self-biasing arrangement, shown in Fig. 162*b*. Here, the bias resistor is connected into the common anode circuit of all the valves and carries the total anode current. In this case, the bias resistor serves as a voltage divider, and the grids requiring various values of voltage are connected to different points on such divider. Of course, if the valves require similar values of bias voltage, their grid circuits are all connected to the negative terminal of the anode power supply. A common self-biasing circuit of such type can be also used with cathode-type valves, although it has an inherent disadvantage, namely that the bias voltage of any valve in the circuit depends

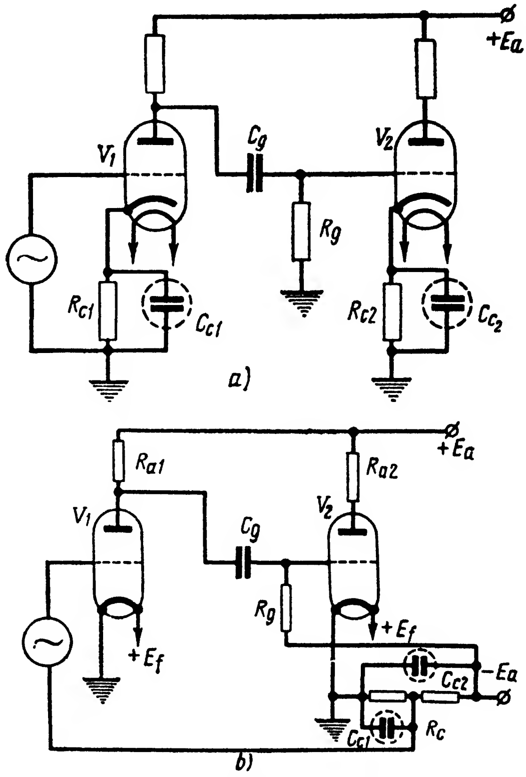


Fig. 162. Automatic bias voltage supply in amplifiers, employing cathode-type valves *(a)* or any type of valves *(b)*

upon the anode currents of all the other valves. However, sometimes this disadvantage of the circuit is turned into its advantage. For instance, in the circuit diagram shown in Fig. 162*a* it is impossible to obtain sufficient bias voltage to cut off the anode current of a valve (a valve is not able to cut off its own anode current); in contrast with this, the circuit diagram shown in Fig. 162*b* offers a possibility of cutting off the anode current of any valve by the biasing voltage developed by the anode current of the other valves.

Anode Decoupling Filters

In multi-stage amplifiers, a parasitic feedback can occur between various stages, such feedback passing through the common anode supply circuits. Fig. 163*a* gives the simplified circuit of a three-stage amplifier. Let us analyse, for example, the influence exerted by the output stage upon the preceding stages. If the anode power supply source had no internal resistance, the alternating anode current of the third stage would flow through the source (it should be kept in mind that the valve itself acts as the generator of such alternating current). This current would not cause any detrimental effect upon the operation of the preceding stages of the amplifier. However, when the anode power-supply source possesses a certain amount of internal resistance (which is always the case), a part of the alternating anode-current of the output stage is branched off

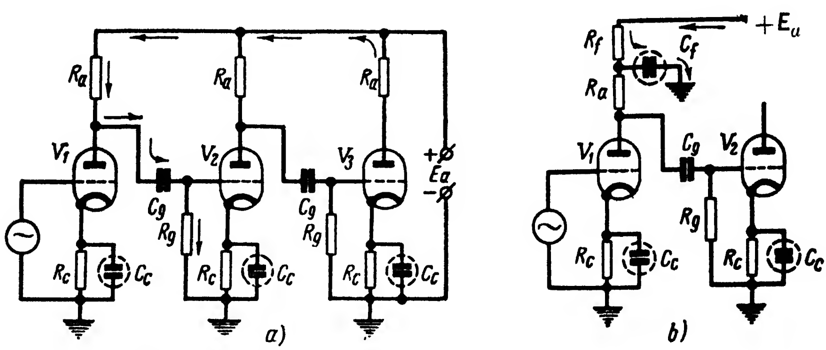


Fig. 163. Parasitic feedback through the common anode supply circuit, and the connection of anode decoupling filter

into the anode circuits of the other stages, passing through their load resistors R_a, coupling capacitors C_g and grid resistors R_g. The path of this current in the first stage is shown by the arrows in Fig. 163*a*. This current builds up alternating voltage across resistor R_g. The voltage is then amplified, reaches the output stage, where it sets up alternating current, a part of which is again branched off

into the preceding stages, creating again an a.c. voltage on the grids of these stages, etc. This cyclic process can give rise to parasitic oscillations, resulting in squeaks, howls, hum or a peculiar type of noise resembling the operation of a motor ("motor-boating" is the common name of the last effect).

Anode decoupling filters are resorted to in order to eliminate such parasitic feedback occurring as a result of undesirable coupling through the common anode supply circuit.

Filters of this type are connected into the anode circuit of each valve, with the exception of the output stage. Fig. 163*b* shows an anode decoupling filter consisting of resistor R_f and capacitor C_f. The resistance value of R_f is 5,000-20,000 ohms (rarely greater than such values), while the capacitance of C_f (usually an electrolytic capacitor) can be from 4 to 10 mfd and even larger.

Resistor R_f opposes the flow of the alternating anode current of the output stage into preceding stage. And although a certain part of this current manages to pass through R_f, it is returned to the cathode of the output stage through filter capacitor C_f—because the latter presents only insignificantly small capacitive reactance even to currents of low frequency. As a result of this, such a small current flows through R_a, C and R_g, that the alternating voltage built up by this current across R_g is quite negligible and does not affect the normal operation of the amplifier. When such anode decoupling filters are incorporated in an amplifier circuit, the amplifier gives a stable and clear performance.

Grid Decoupling Filters

Automatic biasing arrangement, shown in Fig. 162*b*, can also cause parasitic feedback between amplifier stages—the path of such feedback passing through grid circuits. The possibility of such feedback is obvious from the following. R_c carries the alternating current of the output valve, this current setting up a.c. voltage across R_c. This voltage, together with the bias voltage, is applied to valve grids of the preceding stages and is capable of causing parasitic oscillation. Shunting the bias resistor by means of a large capacitor reduces the a.c. voltage across R_c. This, however, is insufficient, particularly on lower frequencies of the audio range, when the capacitive reactance of C_c is not low enough.

Because of this, a decoupling filter C_fR_f (Fig. 164) is connected into the grid circuit of each valve. R_f has a value of several hundred thousands ohms, while the value of C_f is expressed in several tenths of a microfarad. C_f and R_f, connected as shown, form a voltage divider. The capacitive reactance of C_f is many times smaller than the resistance value of R_f.

Because of this, only an insignificant part of alternating voltage, built up across the bias resistor R_c, will appear across C_f. It is this

greatly reduced alternating voltage that is fed to the valve grid from capacitor C_f. Hence, the parasitic coupling is weakened many times, since almost all of the alternating voltage is lost in R_f, practically failing to reach the grid of the given valve.

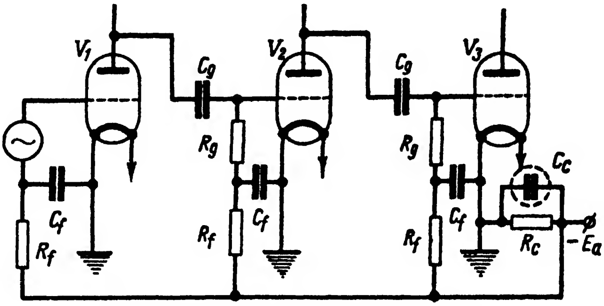


Fig. 164. Grid decoupling filters

When amplifier stages employ individual automatic biasing circuits, as shown in Fig. 162*a*, there is no need for grid decoupling filters, because in such a circuit no interstage feedback can occur through the grid circuits. Still, even such amplifiers sometimes employ a grid filter for suppressing any negative feedback in the stage itself.

Tone Control and Tone Correction

A tone control is often included in amplifiers for the purpose of changing at will the frequency characteristic, thereby also changing the timbre of sounds. Circuit shown in Fig. 165*a* is resorted to when it is desired to reduce the amplification of the higher audio frequencies (i.e., to attenuate these frequencies). The lower the value of resistor R, the greater becomes the shunting effect of capacitor C upon the primary winding of the transformer. The value of C is so selected, that the capacitive reactance of this capacitor is high on the lower and medium frequencies and has no effect on signals of these frequencies. In practical circuits, the value of C is made equal to about 0.005-0.01 mfd and the value of R — to several tens of thousands of ohms. The tone control shown in Fig. 165*b* produces quite the opposite effect — it reduces the amplification of the lower frequencies, i.e., attenuates the bass tones. Choke Ch has a high inductive reactance on the higher and medium frequencies and, therefore, does not affect the frequency characteristic on such signals. The shunting action of the choke upon the primary winding of the transformer begins to tell only on the lower frequencies, becoming more and more pronounced as the value of R is decreased. Both described circuits are combined in the tone control shown in Fig. 165*c* in this circuit arrangement.

Attenuation of either the higher or the lower frequencies can be obtained by moving the potentiometer slider.

When extreme simplicity is required, some amplifiers employ a step-type tone control, instead of the potentiometer (Fig. 165*d*).

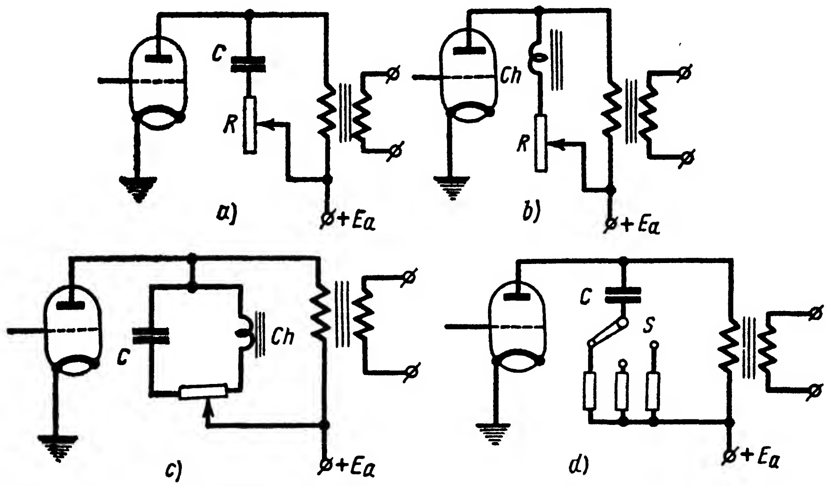


Fig. 165. Various tone-control circuits

Here, a selector switch *S* cuts in at will any one of different fixed resistors, thereby changing the frequency characteristic of the amplifier in "jumps". Some circuits use fixed capacitors instead of the resistors.

Besides tone controls, tone correctors are also frequently used by amplifiers for the purpose of improvement of the amplifier frequency characteristic. For instance, in pentode-output stages the primary winding of the output transformer is sometimes shunted by a capacitor and resistor, connected in series. In this case, the capacitor has a value of several hundredths of a microfarad, while the resistor generally has a value of several tens or hundreds of thousands of ohms. On the higher audio frequencies, the impedance of such a circuit is decreased, which compensates for the increase of the inductive reactance of the primary winding. This reduces the distortion to a considerable degree.

Valves

Voltage amplification stages employ triodes or else low-power high-frequency pentodes, which give a very high performance also in the audio range. The screen grid of such a pentode is usually supplied with a lower voltage than its anode, because there is no

need for large anode currents in these stages. The pentodes are sometimes connected as triodes in such application. Occasionally, a two-stage amplifier circuit is assembled around a single twin-triode (Fig. 166). Valves used in the output stages of amplifiers have already been discussed above.

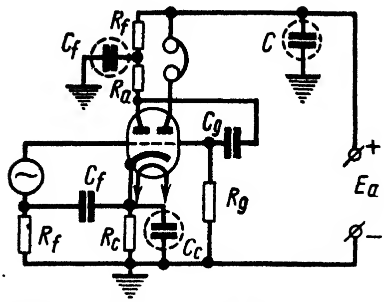


Fig. 166. A two-stage amplifier employing a double triode

79. NEGATIVE FEEDBACK IN AMPLIFIERS

The performance of an amplifier can be improved by including a negative-feedback circuit in the amplifier. In a negative-feedback circuit, a certain part of the amplifier output voltage is fed to the amplifier input circuit, the phase of such voltage being opposite to the phase of the input voltage.

When a negative feedback is used, the improvement of the amplifier performance is attributed to the following:

1) the reduction of frequency distortion and also of non-linear distortion;

2) the reduction of a. c. mains hum;

3) stabilisation of the amplifier gain and a lesser dependence of the gain upon supply-voltage changes, output load-resistance changes, valve replacements, etc.;

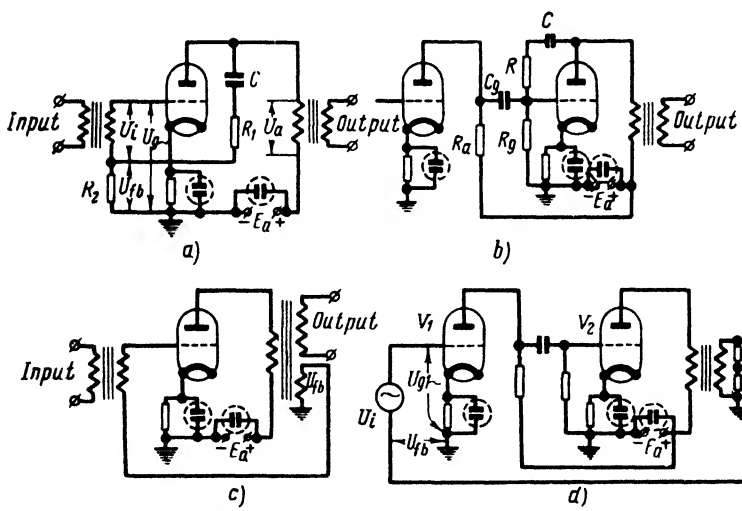


Fig. 167. Negative-feedback circuits

4) considerable reduction of the anode-resistance of the amplifier valve, permitting efficient operation of the output circuit of the amplifier into a low-resistance load;

5) the possibility of controlling the frequency-characteristic of the amplifier, thus effecting the frequency-correction by means of the negative feedback;

It should be noted, however, that not all of the above-listed advantages are realised by every negative-feedback circuit arrangement.

The reduction of non-linear distortion should be considered as the main advantage of negative feedback. Such distortion chiefly arises in the final stage of an amplifier, and, hence, this is the stage which should primarily incorporate a negative-feedback circuit. In some cases, the action of a negative feedback circuit is also extended to the penultimate stage which is used to drive the final stage. The extension of a negative-feedback circuit to a greater number of stages in an amplifier becomes more difficult because there arises a possibility that, at certain frequencies, the negative feedback will be changed to positive feedback and this will cause parasitic oscillation in the amplifier. Such a possibility also exists, although to a lesser degree, when the negative feedback is applied only to two stages of an amplifier. A considerable reduction of the overall gain of an amplifier, when the negative feedback is employed, is a serious shortcoming of all feedback arrangements. This decrease of amplification can be compensated for by an increase of amplification in the first stages of the amplifier, which is quite feasible and does not noticeably increase the distortion, which is very low in such stages, anyway. The reduction of gain because of the feedback is eliminated in specially developed *balanced* negative-feedback circuits.

Let us now discuss the action of a negative-feedback circuit applied to the final stage of an amplifier and shown in Fig. 167*a*. In this circuit, voltage U_a, developed by the primary winding of the output transformer, is applied to a divider consisting of resistors R_1 and R_2. Henceforth we shall be considering only alternating voltages. The value of capacitor C, connected in series with these two resistors, is so selected (0.1-0.5 mfd) that the capacitive reactance of this capacitor is quite low over the whole range of audio frequencies. The purpose served by this capacitor is that of keeping d.c. anode voltage away from resistors R_1 and R_2. The negative feedback voltage U_{fb} is taken from across resistor R_2 and is fed to the grid of the valve. This voltage is in phase opposition to the input voltage U_i, supplied by the preceding stage. The voltage at the grid of the valve is, thus, equal to:

$$U_g = U_i - U_{fb};$$

hence:

$$U_i = U_g + U_{fb},$$

which means that U_i must be greater than U_g.

The amplifier gain, when no feedback is employed, is given by the following:

$$k = \frac{U_a}{U_g};$$

hence:

$$U_a = kU_g.$$

The ratio of negative-feedback voltage U_{fb} to voltage $U_{a\sim}$ is known as the *feedback factor β and indicates what part of the alternating voltage, present in the anode circuit or at the output, is fedback to the grid circuit or to the input of an amplifier.* Mathematically, this is expressed as follows:

$$\beta = \frac{U_{fb}}{U_a};$$

hence:

$$U_{fb} = \beta U_a.$$

In an amplifier, the value of β may be from 0.05 to 0.2.

The amplification factor (the gain) of a stage, when negative feedback is employed, is given by the following:

$$k' = \frac{U_a}{U_i}.$$

But:

$$U_i = U_g + U_{fb} = U_g + \beta U_a = U_g + \beta k U_g = U_g(1 + \beta k).$$

On the basis of the equation, the following expression may be written:

$$k' = \frac{U_a}{U_g(1 + \beta k)};$$

or finally:

$$k' = \frac{k}{1 + \beta k}.$$

Thus, the introduction of feedback into an amplifier circuit reduces the gain of the circuit by $1 + \beta k$ times. Frequency distortion, non-linear distortion, as well as a.c. mains hum can be decreased by the same number of times with the help of negative feedback. For instance, the coefficient of non-linear distortion, when feedback is employed, will be given as follows:

$$K'_n = \frac{K_n}{1 + \beta k},$$

where K_n is the coefficient of non-linear distortion in the absence of feedback.

In the absence of feedback, the input voltage is equal to Ug, while, when the feedback is used, it is given as $U_i = U_g(1 + \beta k)$. Hence, in order to compensate for the reduction of gain by $1 + \beta k$ times, it becomes necessary to apply to the input circuit a voltage which is by $1 + \beta k$ times higher than in the absence of the negative feedback. If this is done, the output power will not be decreased.

Let us now take a numerical example to illustrate how the distortion is reduced in the presence of negative feedback. Assume that an amplifier stage has a $k = 50$ gain on a medium frequency and when no feedback is used. Now, introducing a negative feedback with a factor $\beta = 0.1$, it can be shown that the gain becomes equal to the following figure:

$$k' = \frac{50}{1 + 0.1 \times 50} = \frac{50}{6} \approx 8.3.$$

The gain has decreased by 6 times, because $1 + \beta k = 6$. Now let us assume that, in the absence of negative feedback, $k = 40$ at a certain low or high frequency of the audio range; i.e., an attenuation of 20% is observed. Applying the negative feedback, we find that the attenuation has decreased on this frequency as follows:

$$1 + \beta k = 1 + 0.1 \times 40 = 5;$$

hence:

$$k' = \frac{40}{5} = 8.$$

Comparing this with the amplification on the medium frequency, we see that the attenuation is now expressed only by 4%.

Apparently, the frequency distortion has been decreased by $1 + \beta k$ times (by 5 times, in the given example). The lower the amplification on any frequency, the smaller will be the output voltage. But then the negative feedback voltage will be correspondingly smaller, and, hence, the greater will be the grid voltage, which will to a certain extent compensate for the decrease of amplification on the given frequency. A similar result can be obtained when dealing with an increase of amplification on any frequency. An amplifier incorporating a negative-feedback circuit automatically levels out its frequency characteristic.

The aforesaid discussion applies equally well to changes in the coefficient of amplification which arise from causes other than changes in frequency. Thus, whatever the reason for the change in the coefficient of amplification, the relative (or precentage), this change, with negative feedback, is always $1 + \beta k$ times smaller than a similar change in a non-feedback circuit; or in other words,

the amplification becomes more constant. Mathematically, this can be expressed thus:

$$\frac{\Delta k'}{k'} = \frac{\frac{\Delta k}{k}}{1 + \beta k}.$$

An interesting case of negative feedback is observed when the value of βk is considerably greater than 1. Here, the following expression may be written:

$$k' \approx \frac{1}{\beta}.$$

It so happens, that the amplification does not depend upon k, or upon any other factor, but is entirely determined by the value of β. In such a case, a very stable, although a low, amplification is observed. The amplification practically does not depend upon frequency, which means that frequency distortion is almost absent.

The following example will illustrate the decrease of non-linear distortion in the presence of negative feedback. The curves given in Fig. 168*a* represent a sinusoidal input voltage and distorted output voltage in an amplifier incorporating no negative-feedback circuit (U_i and U_o are drawn to different scales). In the given case, the amplifier develops non-linear distortion of such a character that the first (positive) half-wave of the output voltage has a considerably larger amplitude than the second (negative) half-wave. Fig. 168*b* gives similar curves depicting the operation of the amplifier with negative feedback.

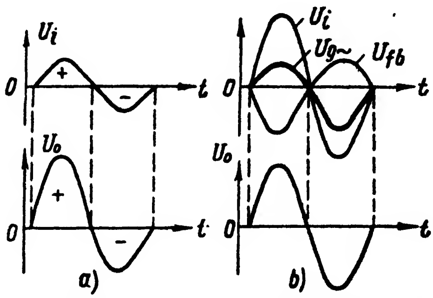


Fig. 168. Reduction of non-linear distortion by means of negative feedback

The input voltage U_i is still sinusoidal. The amplitude of this voltage had to be increased. The negative-feedback voltage U_{fb}, which is in phase opposition to voltage U_i, is noted for the greater amplitude of its first half-wave and for the smaller amplitude of its second half-wave, because it is a part of the output voltage. Grid voltage U_g, equal to the difference between U_i and U_{fb}, is shown by a thick line.

This voltage is noted for the smaller amplitude of its positive half-wave and for the greater amplitude of its negative half-wave. Since the positive half-wave is amplified to a higher level in the given amplifier, the shape of voltage at the output of the amplifier will be nearly sinusoidal, as shown in the drawing. A reduction of non-linear distortion, thus, has actually taken place.

The anode resistance of a valve functioning in an amplifier with negative feedback is decreased by $1 + \beta\mu$ times and is found from the following equation:

$$R_i' = \frac{R_i}{1 + \beta\mu}.$$

The amplification factor of the valve has been reduced by the same number of times

$$\mu' = \frac{\mu}{1 + \beta\mu}.$$

The mutual conductance of the valve remains unchanged. Any valve, in this case, begins to act as a triode and, therefore, the load resistance value R_a must range from $2R_i'$ to $3R_i'$.

Continuing our study of negative-feedback circuits, we note the following. In the circuit shown in Fig. 167*a*, the negative feedback factor is given by the following expression:

$$\beta = \frac{R_2}{R_1 + R_2}.$$

The total resistance $R_1 + R_2$ is made about 20 times as large as R_a.

The negative-feedback circuit studied above is called *the series negative feedback*, while the arrangement shown in Fig. 167*b* is known as a *parallel negative feedback*. In the latter case the following relation holds true:

$$\beta = \frac{R_{com}}{R_{com} + R},$$

where: R_{com} is the total resistance of parallel-connected R_a, R_g, and R_i of the preceding valve.

The disadvantage of these circuits is seen in that they do not reduce the hum generated by anode voltage pulsation. The voltage of such pulsations will reach the anode and grid of the valve (through the negative-feedback divider) in the same phase and will amplify anode-current pulsations. A circuit, free from this shortcoming, is shown in Fig. 167*c*. In this circuit, the negative-feedback voltage is generated by an auxiliary winding of the output transformer. A negative-feedback circuit, extending to two stages, is shown in Fig. 167*d*. Voltage U_{fb} is obtained from a divider provided in the output; hence:

$$\beta = \frac{R_2}{R_1 + R_2}.$$

In this case, we consider that the k value stands for the full amplification factor of the two stages, i.e:

$$k \approx \frac{U_0}{U_{g1}}.$$

The strongest negative feedback is obtained in a *cathode-loaded circuit*, otherwise known as *cathode-output circuit*. In a circuit of this type, shown in Fig. 169, the anode load resistor R_a is connected not into the anode circuit, but into the circuit of the cathode. In this circuit, the whole of the output voltage U_0, built up across the load resistor, is, in effect, a negative-feedback voltage. Accordingly, $\beta = 1$ and $k' = \frac{k}{1 + k'}$, i.e., the value of k' is slightly smaller than unity.

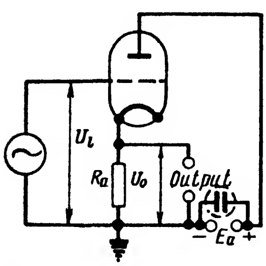


Fig. 169. Circuit diagram of an amplifier with cathode output (cathode follower)

A cathode-output stage produces very small distortion, has a low input capacitance, a high input impedance and a low output resistance. This circuit has found application in final and driver stages of special broad-band amplifiers, designed to give an even-level amplification over a wide frequency band and in some other cases. A cathode-output stage is noted for the following interesting peculiarity: its output voltage U_0 is not only almost equal to the input voltage U_1, but the two voltages also coincide in phase — something, which does not happen in an ordinary amplifier. Because of this, a cathode-output stage is usually referred to as *cathode follower*.

All the circuits studied above employ the so-called voltage feedback. Current feedback is also used in some radio circuits, although to a much smaller extent. A current feedback circuit does not differ from the usual automatic grid bias circuit (Fig. 149*a*). The only difference between the circuits is that the feedback

circuit does not use a capacitor to shunt the cathode resistor R_c. The alternating voltage built up across R_c is the negative feedback voltage. This voltage is proportional to the anode current of the valve. The current feedback decreases non-linear distortion and hum, decreasing at the same time the amplification of the stage. However, this type of negative-feedback circuit does not decrease the frequency distortion and can even aggravate it.

Frequency correction becomes possible if the value of β is made different for various frequencies. To do this, reactances (i.e., capacitors and coils) are included in a negative feedback circuit. For instance, if in the circuits of Figs. 167*a* or 167*d* resistor R_2 is shunted by a capacitor of such a capacitance that its influence would be pronounced only on the higher audio frequencies, the value of β will be reduced and the amplification increased on such frequencies. On the other hand, if such a capacitor is connected across R_1, the value of β will be increased on the higher audio frequencies and a strong attenuation of these frequencies will take place. This principle may be used as a method of tone control.

80. QUESTIONS AND PROBLEMS

1. Given a five-stage amplifier. The amplification factor of each one of the first three stages is equal to 20. In the fourth stage $k_4 = 10$, in the final stage $k_5 = 4$. Determine the amplification factor (the gain) of the whole amplifier.
2. Does the output power of an amplifier depend upon the value of load resistor into which the final stage of the amplifier works?
3. An amplifier gives linear amplification within a frequency range of 300-3,000 cps. Is this a sufficiently good amplifier performance?
4. Plot the frequency characteristic of an amplifier, in which the lower frequencies are emphasised (excessively amplified), while the higher frequencies are attenuated.
5. What is the correction of frequency distortion?
6. The coefficient of non-linear distortion K_n of an amplifier is equal to 7%. What does this mean?
7. What types of distortion are introduced by electron valves and why?
8. Find the gain of an amplifier stage employing a valve with the following parameters: $\mu = 12$; $R_i = 8,000$ ohms, when the load is given by $R_a = 40,000$ ohms.
9. Why is it that the source of anode voltage is not shown in the equivalent circuit of an amplifier stage?
10. Draw an equivalent anode circuit for the d.c. component of the anode current of an amplifier stage.
11. Reproduce the dynamic characteristic of a valve (Fig. 77) on graph paper. Select grid bias voltage and give a graphic representation of the amplification process when an a.c. signal with an amplitude value of $U_{mg} = 3.5$ v is applied to the grid of the valve.
12. Referring to the same characteristic, determine the maximum value of alternating voltage that may be fed to the grid for distortionless amplification, and also determine the value of the bias voltage that must be applied to the grid in such a case.
13. What is the purpose served by the negative grid bias voltage in an amplifier stage?
14. Explain why it is recommended to increase the anode voltage of a valve when the amplification of large alternating voltages is to be obtained with the least possible non-linear distortion.
15. Why is it that the final stage of an amplifier usually does not employ resistance coupling?
16. The amplification factor (the gain) of an amplifier is equal to 1,000,000. Will such an amplifier develop an output voltage of 1,000,000 v if an input signal of one volt is applied to its grid circuit?

17. The voltage of anode power supply, feeding a resistance-coupled amplifier, is equal to 350 v. The amplifier draws $I_a = 2.5$ ma and the resistance of its load R_a is equal to 80,000 ohms. Find the anode voltage of the amplifier.

18. A 1,000-ohm resistor is connected to an alternator through a step-down transformer with a transformation ratio of 5:1. What is the value of resistance presented to the alternator?

19. The internal resistance of an alternator is 2,000 ohms. The alternator supplies power to a 500-ohm resistor through a transformer. What should be the transformation ratio of the transformer if the generator is to face a load of 8,000 ohms?

20. In a transformer-coupled amplifier, the inductance of working turns L_1 of the primary winding is 32 h, while the leakage inductance L_l is 0.5 h. Wha is the inductive reactances of these inductances on 100 cps and on 5,000 cps?

21. What are the advantages and disadvantages of a transformer-coupled amplifier, as compared to a choke-coupled amplifier?

22. Why is it undesirable to design an interstage transformer with too small a core?

23. Draw equivalent circuits of a transformer-coupled amplifier, employing parallel anode feed, for the lower and the higher audio frequencies.

24. The d.c. component of anode current I_a of a valve is equal to 2.5 ma. What should be the resistance value of self-biasing resistor R_c, connected into the cathode wire, if it is desired to obtain a bias voltage of −4 v?

25. Is it possible to cut off the anode current of a valve by means of automatic biasing circuit shown in Fig. 149?

26. What is the influence exerted upon the frequency characteristic of an amplifier by capacitor C_c shunting the self-biasing resistor R_c in an amplifier stage? Explain in detail.

27. The final stage of an amplifier employs a triode valve, in which the value of R_i is equal to 1,000 ohms. Would it be a good practice to make such a stage operate into a load of 200 ohms?

28. What should be the load resistance of a final stage, using a pentode valve, the internal resistance R_i of which is equal to 60,000 ohms?

29. Why is it that an output transformer always has a considerably larger core than an interstage transformer?

30. Why is it considered undesirable when the d.c. components of the anode currents of the valves in a push-pull stage have different values?

31. Given a single-ended and a push-pull amplifier stage of similar power output. In which one of these stages the core of the output transformer can afford to have a smaller cross-section?

32. Draw the circuit diagram of a push-pull output stage employing pentodes, each valve provided with its own automatic biasing circuit.

33. What are the advantages of a class A amplifier, in comparison with amplifier classes B and AB?

34. Draw the circuit diagram of a resistance-coupled amplifier, employing a two-section anode decoupling filter.

35. Anode current I_a of a valve is 30 ma. A 1,200-ohm resistor is connected in the anode circuit of this valve to obtain automatic bias voltage. Find the value of the bias voltage.

36. Why cannot be the voice coil of a dynamic loudspeaker connected directly into the anode circuit of a final stage?

37. What is the purpose served by tone controls?

38. Devise the circuit of an amplifier, employing a 6ЖK7 pentode in the first stage, a 6C5 triode in the second stage, and a 6H7 valve in the third (push-pull) stage. A microphone-pickup change-over switch and a gain control must be provided in the input circuit of the amplifier, while a tone control is to be connected in the final stage. All the stages of the amplifier are to use automatic biasing circuits. Anode decoupling filters are to be incorporated in the first two stages. Besides the amplifier circuit, draw the circuit of its associated rectifier

power supply employing a 5Ц4С valve. Show approximate ratings of varions components used in both circuits.

39. A resistance-coupled amplifier stage employs a triode, consumes $I_a =$ $= 2$ ma, works into a load $R_a = 80{,}000$ ohms, and uses a 2,000-ohm self-biasing resistor. The voltage of the anode power supply, used with this amplifier, is equal to 240 v, the negative terminal of the power supply being connected to the chassis. Find the values of anode and grid bias voltages. Also determine the potentials of the following points of the circuit in respect to the chassis: potentials of anode, control grid, cathode, the positive terminal of the anode power supply.

40. A sinusoidal voltage is amplified by an amplifier. In the output circuit of the amplifier, besides the fundamental frequency voltage with an amplitude of $U_{m1} = 20$ v, is also obtained the second-harmonic voltage with an amplitude of $U_{m2} = 1.6$ v. Determine the value of the non-linear distortion coefficient.

41. When it is desired to obtain a high value of inductive reactance of the primary winding of an interstage transformer, why is it customary to avoid the practice of winding such a coil with a very large number of turns, say, a few tens of thousands?

42. An amplifier develops an output power of 70 watts in a load resistor of 140 ohms. Determine the voltage amplitude at the output of the amplifier under such operating condition.

43. On what frequencies does the distributed capacitance of transformer windings introduce distortion?

44. Devise the circuit diagram of a three-stage amplifier employing filamentary valves and designed to operate from a microphone. The first two stages are to use resistance-coupled pentodes. The last stage is also to employ a pentode and to be provided with a transformer output. The amplifier circuit is to be automatically biased by the anode current, the control grids of the first two stages fed with −3 v, while the control grid of the final stage with −10 v of negative bias. Calculate the resistance value of the automatic biasing circuit if the d.c. components of the cathode currents of the valves have the following respective values: $I_{a1} = I_{a2} = 1.5$ ma; $I_{a3} = 22$ ma. Be sure to incorporate anode and grid decoupling filters in the first two stages of the amplifier.

45. The anodes of final stage valves, provided with a fixed bias, are heated to dull-red colour during the transmission of speech but show no signs of heat during pauses. Does such final stage operate as a class A or class AB amplifier?

46. The resistor used in the automatic biasing circuit of an output stage has a resistance of 200 ohms. It is required that the capacitive reactance of the capacitor shunting such resistor is ten times as small as the resistance value of the resistor on the lowest frequency being transmitted, namely—on 50 cps. What is the capacitance value of such a capacitor?

47. What purpose is served by phase-inverting circuits in amplifiers?

48. What is a negative-feedback circuit?

49. What advantages does a negative-feedback circuit offer to an amplifier?

50. If the negative-feedback factor $\beta = 0.2$, by how many times will be decreased the gain of an amplifier if its k was equal to 100 before the negative-feedback circuit was incorporated?

51. What are the peculiarities of a cathode-output amplifier stage?

CHAPTER VIII

VALVE OSCILLATORS AND TRANSMITTERS

81. SELF-EXCITED VALVE OSCILLATORS

Valve oscillators are employed to generate alternating currents of various frequencies. The most important of such oscillators are the high-frequency oscillators used by radio transmitters, radio receivers and radio measuring apparatus. Audio-frequency oscillators are also employed by radio equipment, although not as often as high-frequency oscillators.

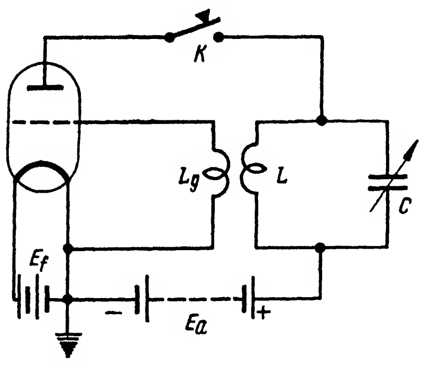


Fig. 170. The circuit of a valve oscillator with inductive feedback

One of the simplest types of valve oscillators was discussed in Chapter IV, Sec. 33. It consists of an electron valve (a triode or a more complex valve), a tuned circuit, power supplies and a feedback circuit (Fig. 170).

It is a customary practice to group all valve oscillators into the following two classes:

1) *separately-excited oscillators, in which the grids are supplied with a.c. excitation voltage from another oscillator, commonly called an exciter;*

2) *self-excited oscillators in which the grids are supplied with a. c. excitation voltage from their own tuned circuits.*

The oscillator shown in Fig. 170 and employing inductive feedback belongs to the class of self-excited oscillators. Generally speaking, the presence of a feedback circuit in an oscillator is an indication that the oscillator belongs to the self-excited class.

In the circuit shown in Fig. 170, anode current flow begins when key K is closed. This current charges the capacitor of the tuned circuit, thus commencing free damped oscillations in the given circuit. Alternating current flowing through coil L induces a.c.

voltage in grid coil L_g; this voltage, being applied to the grid, causes anode current pulsations. An a.c. component appears in the anode current. The valve itself acts as the generator of such a.c. component, just as it does in any amplifier stage. The a.c. component of the anode current passes through tuned circuit LC, generating a. c. voltage in this circuit. This voltage is, in effect, an amplified a.c. grid voltage, the amplification being provided by the valve, the tuned circuit acting as the latter's load impedance. The frequency of a.c. grid voltage is equal to the frequency of natural oscillations of the tuned circuit. Hence, the a.c. component of the anode current has the same frequency. This means that the state of parallel resonance always exists in the tuned circuit connected to the anode, this tuned circuit presenting high impedance to the a.c. component of the anode current.

The oscillations set up in the tuned circuit upon closing of the key will soon damp out, unless they are sustained by the a.c. component of the anode current and converted into continuous (undamped) oscillations. To do this, it is mandatory that the phase of the amplified voltage, built up across tuned-circuit LC by the a.c. component of the anode current, coincides with the phase of the free oscillation voltage in the tuned circuit. If the two voltages fall out of phase, the commenced oscillations are sure to damp out quickly and the circuit will fail to excite itself.

Such state of affairs can be best illustrated by drawing an analogy to the classical theory of mechanical pendulum. If a heavy pendulum is made to swing — (oscillate) — and if its oscillatory state is to be sustained, the pendulum must be given a push after regular periods of time. It is not only necessary that the periodicity — (the frequency) — of such pushes coincides with the natural frequency of the pendulum, but also necessary that the phase of the external pushing force coincides with the phase of pendulum oscillations. If the required phase-coincidence is not observed, for instance, if the pendulum is pushed in the direction opposite to its travel, the pendulum will stop swinging after a short lapse of time.

In the electrical oscillator the correct feedback-phase-relation is set by the proper connection of the ends of coils L and L_g. In practice, when an inductive-feedback oscillator fails to excite itself, it is sufficient to reverse the ends of coil L_g and then, as a rule, the oscillator will begin to function (unless some other type of fault is present in the circuit). When the oscillator coils are connected correctly, the alternating voltages appearing at the grid and the anode of the valve will be in phase opposition. This is easy to understand, taking into consideration the following. When the tuned circuit of the oscillator first began to oscillate, a loss of energy took place in the tuned circuit resistance during the first quarter of the period when the capacitor was discharged through the coil. This energy loss must be compensated for by the a. c. component of the

anode current of the valve during the next quarter of the period, when the capacitor is charged by the self-inductance e.m.f. of the coil.

If, for instance, during this quarter of the period the capacitor plate connected to the anode (the upper plate in Fig. 170) is charged negatively (i.e., if the polarity of the alternating anode voltage is negative at the moment), the positive half-wave of the anode current must occur; in other words, the current must increase so that new electrons would reach the upper plate of the capacitor and bring up the capacitor voltage to the former value. But the anode current increase will be caused only when a positive half-wave of the alternating voltage is applied to the grid of the valve. It, therefore, becomes apparent that the voltage at the anode of a properly functioning oscillator valve is opposite in phase to the voltage applied to the grid of the valve.

Correct operation of an oscillator calls for correct phase relationship, as shown above. It also calls for sufficient value of feedback. If the feedback value is too low, the alternating grid voltage will generate only a weak a.c. component of the anode current, and the energy of such component will not be high enough to compensate for the losses in the tuned circuit.

Thus, the conditions necessary for the proper operation of a self-operation of a self-excited oscillator are summed up as follows:

1) *alternating anode and grid voltages must be 180° out of phase;*

2) *the value of feedback must be sufficient.*

As far as its operating principle is concerned, a self-excited oscillator differs but slightly from an amplifier stage. The oscillations generated in the tuned circuit are fed to the grid of the valve via the feedback circuit; the amplified alternating voltage, which is built up across the tuned circuit, is again applied to the valve grid via the feedback circuit, and is again amplified, etc. The amplitude of oscillations is gradually increased, finally reaching a certain limit. As may be seen, a self-excited valve generator acts as an amplifier which amplifies its own oscillations.

82. OPERATING CONDITIONS, POWER AND EFFICIENCY OF A VALVE OSCILLATOR

If the feedback is intensified in a valve oscillator, for instance, by bringing closer together the two coils L and L_g in the inductive-feedback circuit, the alternating grid voltage will increase, the a.c. component of the anode current will also increase, and the oscillations in the tuned circuit will become stronger. This will be carried on up to a certain limit and no further, because the anode current will reach its saturation point on high values of grid voltage and, besides, the considerable grid current that will begin to flow will

consume a part of the oscillatory energy of the tuned circuit. Such an operating condition of an oscillator is called the *overexcited condition*.

The other extreme — when the oscillator functions with a very small value of feedback — is called the *underexcited condition*. When an oscillator is adjusted to operate between these two extreme conditions, it is said that its operating condition is *normal*. This normal condition is attained at a certain average value of feedback. Under the normal operating condition the anode current usually changes from zero to the saturation value, but the grid current is not excessive. Fig. 171 pictures all the three operating conditions.

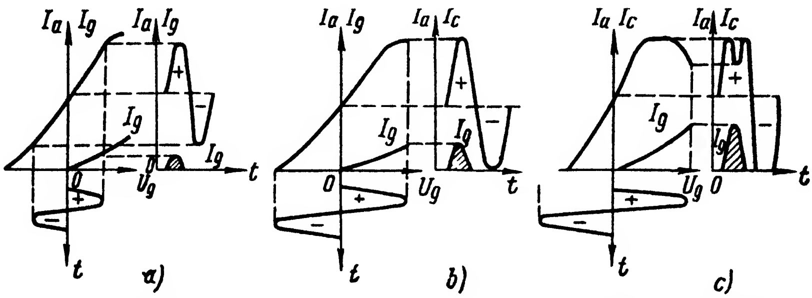


Fig. 171. Underexcited *(a)*, normal *(b)* and overexcited *(c)* operating conditions of a valve oscillator

A steady increase of signal voltage applied to the grid of a valve first changes the underexcited operating condition of the valve (Fig. 171*a*) to the normal condition (Fig. 171*b*) and then to the overexcited condition (Fig. 171*c*).

The last one of the named conditions is noted for a distortion of the sinusoidal shape of anode-current oscillations, "depressions" being formed in the anode current because of a large grid current flow.

The following types of power are considered in the study of valve oscillator circuits:

1) *input power P, provided by the anode supply source and consumed by the anode circuit of the oscillator;*

2) *anode dissipation power, or simply anode dissipation, denoted by P_a and representing the power used to heat the anode;*

3) *oscillatory or useful power P_k developed in the tuned circuit of the oscillator.*

The input power is consumed to provide both the useful power and the anode dissipation power. Hence, the following relation may be written:

$$P = P_k + P_a.$$

If an oscillator does not function, $P_k = 0$ and all of the input power is wasted as anode dissipation, i.e., $P = P_a$. This is the most

undesirable condition in an oscillator, for such a condition can lead to overheating and even melting of the valve anode. This is why stopping of oscillations must be prevented in a functioning oscillator, if the oscillator valve is to be protected from damage.

The value of the useful power that a valve oscillator can give is computed as follows:

$$P_k \approx 0.2\, I_{sat} U_a,$$

where: I_{sat} is the saturation current and U_a is the anode voltage of the valve.

As seen from this formula, the useful power can be increased by raising the anode voltage. Depending upon the operating condition, the actual power developed in the tuned circuit can sometimes be less than the power computed by the formula. But in any case the anode dissipation power must not exceed the maximum permissible value $P_{a\ max}$.

The *efficiency* of a valve oscillator is an important index of its operating condition. The efficiency, in this case, *is the ratio of the useful power to the input power; the efficiency shows what part of the input power is turned into the useful power.*

The efficiency is usually expressed in per cent, which necessitates multiplication of P_k and P by 100. Thus:

$$\eta\% = \frac{P_k}{P}\, 100.$$

For instance, if $P = 10$ w and $P_k = 6$ w, $\eta = \frac{6 \times 100}{10} = 60\%$. The efficiency of valve oscillators can be as high as 70-80%; while low-power oscillators have smaller efficiency values.

The input power can be calculated when the d.c. component I_{a-} of the anode current and the voltage U_a of the power supply source are known. In a functioning oscillator these values are measured with a milliammeter and voltmeter. The formula applicable in this case is as follows:

$$P = I_{a=} U_a.$$

If, for instance, $U_a = 400$ v and $I_{a=} = 50$ ma $= 0.05$ a, the value of P will be:

$$P = 0.05 \times 400 = 20 \text{ w}.$$

The useful power value depends upon the a.c. component of the anode current. If both the P_k and the efficiency values are to be increased, the a.c. component of the anode current should be made larger and the d.c. component of the same current decreased as much as possible. Utilisation of the whole of the valve characteristic, up to the point of saturation, gives the required increase of the a.c. component. Application of a negative grid bias to the valve grid reduces the d.c. component. Operating conditions shown in

Fig. 171 are the conditions obtained in the absence of bias voltage on the grid of the oscillator valve. In this case, the oscillations are sinusoidal for both the underexcited and normal conditions, and the stage is said to operate under *class A amplifier oscillating condition.* Such oscillating condition is noted for low efficiency (not over 40-45%, and practically even less) and is resorted to only in such low-power oscillators which must develop sinusoidal oscillations. Incidentally, this operating condition is similar to the operating condition of class A amplifiers, discussed in Chapter VII.

Much more advantageous is the so-called *class B amplifier oscillating condition* in which the anode-current swings reach cut-off value. Such operating condition is set by shifting the operating point to the left (by means of proper grid bias voltage E_b) and by increasing the alternating voltage fed to the control grid (i.e., by increasing the excitation). An example of class B amplifier oscillating condition is shown in Fig. 172. This condition, although it pertains to an oscillator, resembles the operating condition of a class B amplifier. Under such condition, the shape of anode current pulses depends upon the bias voltage value and also upon the value of the exciting voltage. Since the anode current pulses are separated by time intervals, during which the anode current is absent, the d.c. component is smaller in this type of oscillator than in an oscillator working under class A operating condition.

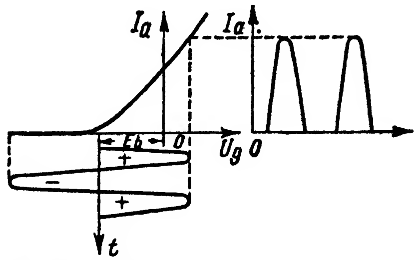


Fig. 172. Class B amplifier oscillating condition

The longer the time intervals between the pulses, the smaller is the d.c. component and the higher is the efficiency. The selection of proper values of bias and excitation makes an oscillator stage give the maximum obtainable useful power output at maximum efficiency. A suitable measuring instrument or indicator of high-frequency oscillations, for example, an incandescent lamp (see Fig. 186*a*) or a thermo-ammeter coupled to the circuit, shows the value of useful power developed in the tuned circuit. The minimum value of the input power is indicated by a dip of the anode milliammeter pointer or by the least heating of the valve anode. The bias voltage value securing the most advantageous operating condition can be approximately determined from the characteristic of a valve. The operating point must be located in the very beginning of the characteristic, similarly to the location of the operating point in the case of a class B amplifier. In the case of an oscillator, the point can be located even farther to the left, in the region where the anode current is zero.

Still another operating condition of an oscillator is possible. Under this operating condition, called class C amplifier oscillating condition, anode current pulses last even less than half-period, which distinguishes a class C oscillator from class B oscillators.

In further discussion and in accordance with the radio engineering practice, we shall use the following terminology when speaking of valve-oscillators working under specific operating conditions:

class A oscillator — an oscillator working under class A amplifier oscillating condition, as defined above;

class B oscillator — an oscillator working under class B amplifier oscillating condition, as defined above;

class C oscillator — an oscillator in which the grid bias is appreciably greater than the cut-off value so that the anode current in the valve is zero when no alternating grid voltage is applied; the anode current in the valve flows for appreciably less than one half of each cycle when excitation is applied to the grid circuit of the oscillator valve.

The most popular method of biasing valve oscillators is that of employing a grid leak. In this method shown in Fig. 173 *a* and *b*, the bias voltage is developed by the flow of grid current. The excitation is applied to the grid through grid capacitor C_g (several hundred picofarads), while the d.c. component of the anode current flows through grid-leak resistance R_g, developing a voltage drop across it, this voltage drop being the required bias. Fig. 173 shows the direction of grid-current flow and the polarity of voltage at the ends of grid-leak R_g. The value of R_g ranges from several thousand to several tens of thousands of ohms.

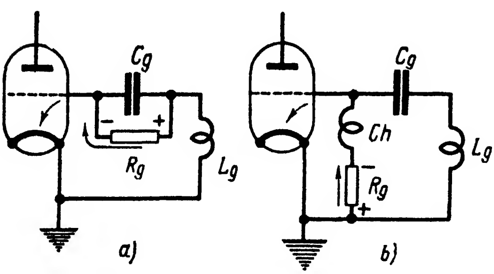


Fig. 173. Automatic building up of bias voltage by means of a grid leak

Grid-leak bias is particularly suitable for application in self-excited oscillators. When a generator provided with grid-leak bias is not functioning, there is no grid current, and consequently, no bias voltage. The operating point, in this case, is located on the linear part of the characteristic. This part of the characteristic has the steepest slope, which is a necessary condition for easy commencement of oscillations. As soon as the stage begins to oscillate, the positive half-waves of the alternating voltage, applied to the grid, give rise to the grid current. The latter generates the bias voltage, making the operating point shift to the left — just what is needed

for the operation of a class B or class C oscillator, working under the condition of anode current cut-off. Application of grid-leak bias secures the most stable operation of a self-excited oscillator. In the circuit shown in Fig. 173*b*, the grid leak (R_g) is connected between the grid and cathode of the valve. In order to decrease the loss of high-frequency power in R_g, choke *Ch* is connected in series with this resistor.

Separately-excited oscillators sometimes employ bias voltage generated by an external source. They also employ automatic biasing circuits, where the bias voltage is developed by the flow of anode current through a self-biasing resistor. The peculiarities of these two biasing methods were discussed in detail in the chapter dealing with low-frequency amplifiers (see Sec. 75).

83. SELF-EXCITED VALVE OSCILLATOR CIRCUITS

Self-excited valve oscillator circuits differ from one another in the methods of feeding a part of the oscillatory anode voltage to the grid so as to sustain the circuit in oscillation. Inductive, capacitive and autotransformer types of feedback are generally used. Another point of difference between various types of oscillators is the method of feeding high d.c. voltage to the anodes of oscillator valves. Here, two such methods are possible: *series anode feed*, when the valve anode and the tuned circuit are connected in series, and *parallel anode feed*, when the valve and the tuned circuit are connected to the anode supply in parallel. And, finally, there is still another point of difference between various oscillators: some oscillators employ push-pull circuits (with two valves), while others use single-ended circuits built around a single valve.

An example of series anode feed is shown in Fig. 170. In circuits employing this type of feed, the tuned circuit is at a high d. c. potential in respect to earth, which is hazardous to the technical personnel servicing such an oscillator. A safer arrangement is shown in Fig. 174*a*. Here also, the oscillator employs an inductive feedback but uses a parallel anode feed, whereby the a. c. and d. c. currents are separated by means of a choke and blocking capacitor C_a. The blocking capacitor passes the high-frequency current to the tuned circuit from the valve anode, but isolates the high d. c. voltage of the anode power supply from coil L. The capacitance of C_a is quite large (from several hundred to several thousand picofarads). To avoid the puncture of this capacitor by high voltage, the capacitor must be rated at double the anode voltage. The d. c. component of the anode current easily passes through the choke *Ch*, because the resistance of this choke is negligible. But the choke has a considerable inductive reactance, and, hence, blocks the high-frequency current, keeping it away from the anode power

supply. If no choke were employed in the circuit, the circuit would not oscillate because the a. c. component of the anode current would not flow through the tuned circuit but would be short-circuited through the anode power supply, the latter possessing a much lower resistance in comparison with the impedance of the tuned circuit.

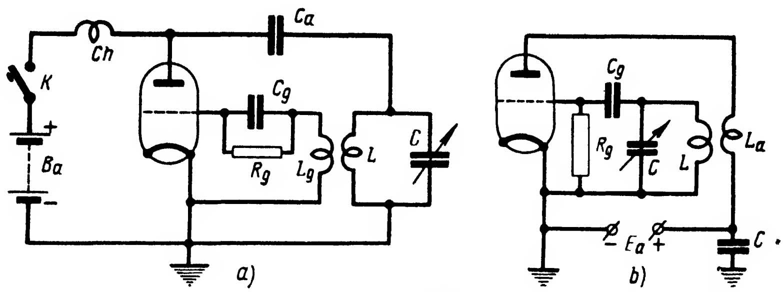


Fig. 174. Various circuits of oscillators with inductive feedback: *a)* circuit with parallel anode feed; *b)* oscillator employing a tuned grid circuit

The presence of the auxiliary components — the blocking capacitor and the anode choke — is a shortcoming (although not a very serious one) of the parallel anode feed circuit.

Low-power oscillators sometimes also employ inductive-feedback circuits, in which the tuned circuit is connected into the control grid circuit, rather than into the anode circuit (Fig. 174*b*).

Fig. 175*a* shows an oscillator employing an *autotransformer feedback* circuit. In this circuit, also known as *tapped-coil* oscillator circuit, the tuned-circuit coil L and the grid-coil represent a single winding. The tuned circuit utilises the whole of coil L, while a part of this coil is the grid-coil section. Thus, the grid is fed with a part of the alternating voltage developed by the tuned circuit, this voltage being delivered by the coil acting as an autotransformer. This circuit could be called a three-point circuit because three points of the tuned circuit participate in the oscillation-generating process: two points—the two ends of the coil connected, respectively, to grid and anode, and the third point—the tap, connected to the cathode. The position of point c (the cathode) between point a (the anode) and point g (the grid) gives a 180° phase shift between the alternating voltages of anode and grid, thus providing the condition necessary for self-excitation of the oscillator. When the oscillator circuit is being adjusted, the proper value of feedback is set by finding the correct position of point c, which is accomplished by the "try-and-cut" method consisting in attaching the cathode wire to various turns of the coil. The optimum position of point c found, the cathode wire is permanently soldered to this point.

In the series anode-feed variety of a tapped-coil oscillator circuit (Fig. 175*b*), grid resistor R_g must not be connected across capacitor C_g, as this would apply the positive high-voltage potential to the grid of the valve through R_g. Fig. 175*b* shows the anode supply source connected between the tuned circuit and the cathode. However, this supply source can be alternatively connected between the tuned

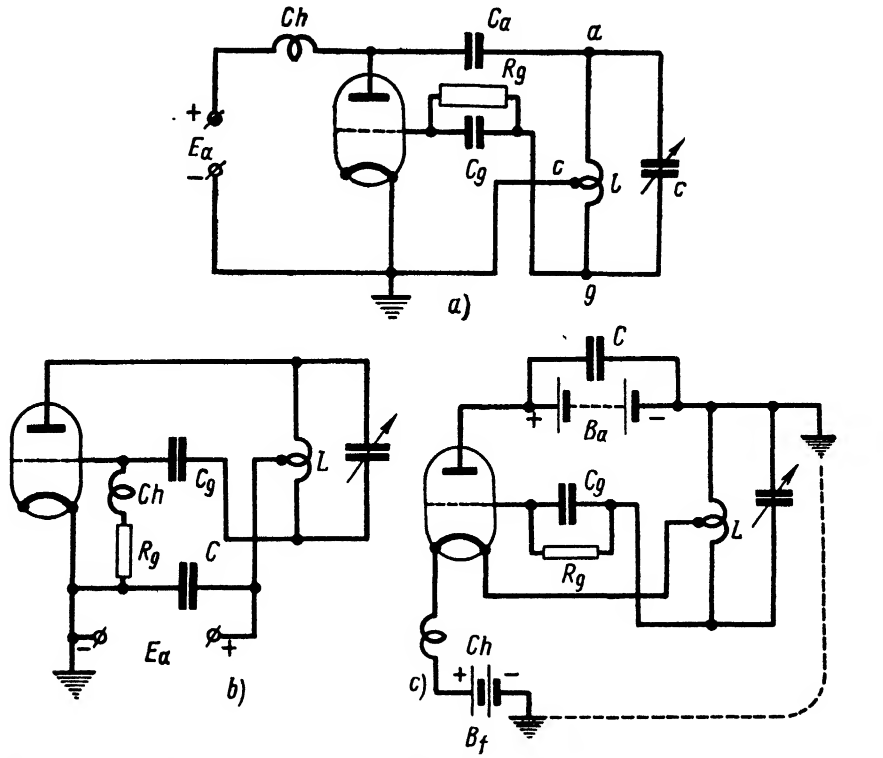


Fig. 175. Various circuits of oscillators with autotransformer feedback: *a)* circuit with parallel anode feed; *b) and c)* circuits with series anode feed

circuit and the anode, as shown in Fig. 175*c*. In the latter circuit arrangement, the chassis is the common negative and is connected to the anode end of the tuned circuit, owing to which the filament current flows through a part of turns of the tuned-circuit coil. Such circuit is known as a *cathode-coupled circuit*, alternatively called an *earthed-anode circuit*, in which the anode is earthed as far as high frequency is concerned.

In a cathode-coupled circuit, the cathode (or the filament) must not be connected to the chassis because the capacitance existing between the anode voltage source and the chassis will shunt a

part of the tuned circuit coil, thus affecting the frequency. When a filamentary valve is used in this type of circuit, a high-frequency choke must be connected into the positive filament wire. If the choke is omitted, the a. c. component of the anode current would not be flowing through the tuned circuit but would be short-circuited through the filament supply, as shown by the broken line in Fig. 175*c*.

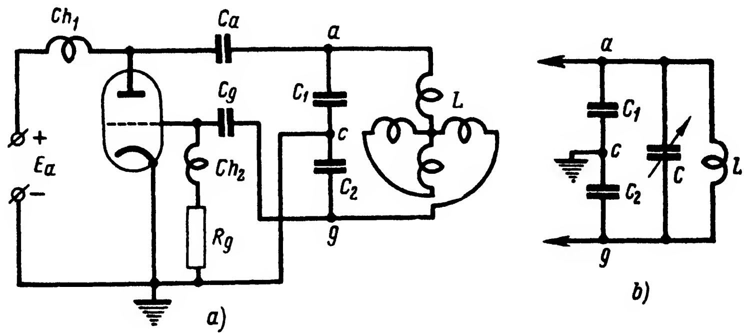


Fig. 176. Oscillator circuits employing a capacitive feedback and parallel anode feed: *a)* variometer-tuned circuit; *b)* capacitor-tuned circuit

When an indirectly-heated valve is used in this circuit, the choke becomes unnecessary because the cathode circuit can be completely isolated from the heater.

The circuit of an oscillator with a *capacitive feedback* and parallel anode feed is shown in Fig. 176*a*. Capacitive-feedback circuits do not use series anode feed, as a rule, and could be also considered as "three-point" circuits. In contrast to an inductive-feedback circuit, the three points of a capacitive-feedback circuit are located on the capacitive branch of the tuned circuit. In this case, the capacitance of the tuned circuit is made up of two series-connected capacitors C_1 and C_2, which form a capacitive voltage divider. Capacitor C_2 is the feedback capacitor, and the voltage developed across it is fed to the grid of the oscillator valve. The oscillatory circuit is tuned by a variometer. When the oscillatory circuit uses a fixed coil. the tuning is provided by a variable capacitor C (Fig. 176*b*).

In a capacitive-feedback oscillator circuit, resistor R_g must be connected between the grid and cathode. Capacitor C_g shown in this type of circuit (Fig. 176*a*) can be actually omitted because the d.c. component of the grid current can not pass through capacitors C_1 and C_2, anyway.

Of special interest is a self-excited oscillator circuit employing two tuned circuits and a series anode feed (Fig. 177). In such an oscillator, the two tuned circuits, adjusted to the same frequency, are placed into the grid and anode circuits, respectively. The feed-

back is provided by the anode-grid capacitance C_{ag} inside the valve, this capacitance no longer playing its usual parasitic role, but a valuable role of a coupling capacitor.

The value of C_{ag} capacitance is small and, therefore, sufficient feedback can be obtained with it only when the oscillator operates on the shorter waves. When it is required that this type of circuit be used on lower frequencies, it becomes necessary to increase the value of the coupling capacitance by connecting an external capacitor between the anode and the grid of the valve, i. e., in parallel with the interelectrode capacitance C_{ag}. This type of circuit, known as *tuned-anode tuned-grid circuit (TATG)*, generally employs a series anode feed, as noted above, but can be alternatively arranged to operate with parallel anode feed.

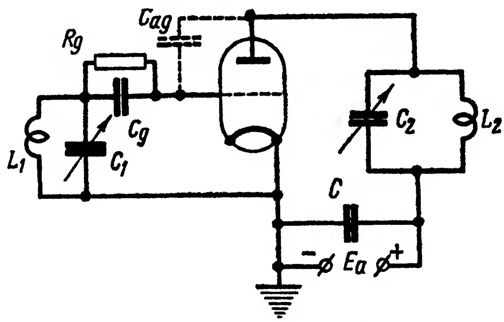


Fig. 177. The circuit of a tuned anode-tuned grid valve oscillator employing a feedback through the anode-grid interelectrode capacitance

All the above-described oscillator circuits are single-ended. However, they can be converted into push-pull circuits employing either series or parallel anode feed. An example of such push-pull circuits is given in Fig. 178, where Fig. 178*a* shows a push-pull

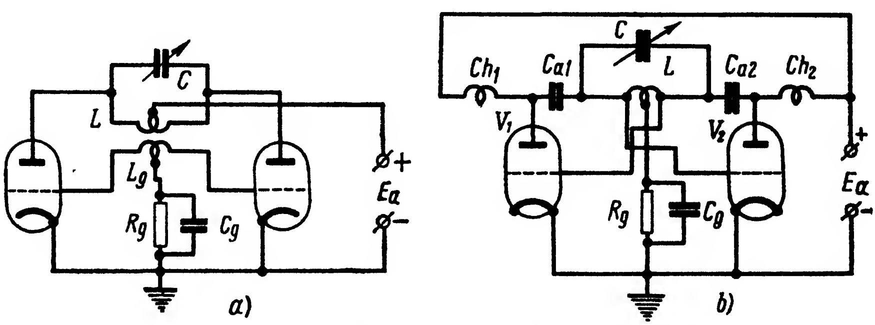


Fig. 178. Push-pull valve-oscillator circuits: *a)* a circuit employing inductive feedback and series anode feed; *b)* a circuit with autotransformer feedback and parallel anode feed

circuit with inductive feedback and series feed, while Fig. 178*b* shows a push-pull circuit with parallel anode feed. The operating principle of push-pull oscillators is similar to the operating principle of low-frequency push-pull amplifiers. As in the case of push-pull

amplifiers, here, too, we see that a push-pull oscillator is comprised of two single-ended circuits employing a common tuned circuit, common power supply and a common grid resistor R_g. The valves operate with a 180° phase shift, i. e., first one valve, then the other.

A push-pull circuit offers the following advantages. It doubles the power output. Its common supply wires do not carry the alternating component of the fundamental frequency (first harmonic) nor the odd harmonics. The tuned circuit of a push-pull stage is free of even harmonics, which are very undesirable in transmitters (see Sec. 85). In a push-pull circuit the interelectrode capacitances of the valves are connected in series and their total value is, thereby, halved. Hence, these capacitances exert smaller influence upon the frequency of the tuned circuit and upon the operation of the stage in general, which makes the push-pull circuit arrangement particularly suitable for operation on short waves, and particularly on ultra-short waves. All the above-named advantages of a push-pull circuit are attained only when there is no asymmetry in the circuit.

84. SELF-EXCITED OSCILLATORS EMPLOYING NO FEEDBACK

Some types of self-excited valve oscillators operate without feedback, by virtue of the dynatron effect. Fig. 179*a* shows the circuit of a *dynatron oscillator*, in which the oscillations are generated by the dynatron action in a tetrode. A tuned circuit is connected into the anode wire of such a tetrode and the anode is fed with lower voltage than the screen grid.

Electrode voltages are so adjusted that the valve operates over the dropping part of the anode characteristic, where an increase of anode voltage results in a decrease of the anode current and vice versa (Fig. 89). The oscillator will sustain oscillations under such operating condition, the oscillations commencing in the tuned circuit as a result of any electrical pulse, for instance—a pulse set up by switching on the anode supply.

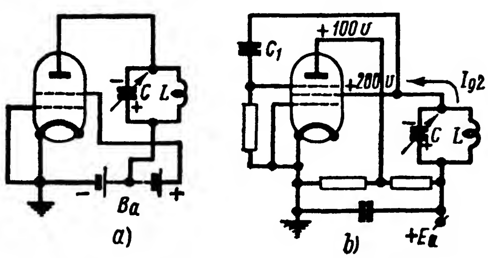


Fig. 179. Dynatron *(a)* and transitron *(b)* oscillators

The dynatron oscillator functions as follows. Assume that the capacitor plate connected with the anode of the valve is charged negatively at a certain moment of an alternating current cycle in the tuned circuit (Fig. 179*a*). In this case, the anode voltage is decreased but the anode current is increased. The increase of the anode current will cause an inflow of a certain quantity of electrons

to the upper plate of the capacitor. Thus, a recharging of the capacitor will take place, and if the capacitor is recharged to a sufficiently high value, the energy losses in the tuned circuit will be compensated for and the oscillations will become continuous. The absence of a special feedback arrangement simplifies the circuit and a "two-point" oscillatory circuit becomes applicable.

Dynatron oscillators are designed only for very low power levels and are mainly used in measuring equipment. These generators provide a stable frequency output but their disadvantage is seen in a certain instability of the secondary emission process.

Transitron oscillators (Fig. 179*b*) are also used as beat-frequency oscillators in certain measuring and radio receiving equipment. In a transitron oscillator, the anode of the valve is supplied with a lower voltage than the screen grid, but the dynatron action does not take place because the valve is a pentode. The tuned circuit is connected into the screen-grid circuit. The suppressor-grid is connected to the cathode through a high resistance and is fed with alternating voltage from the tuned circuit through capacitor C_1 (of the same strength and phase as the voltage on the screen grid). The change of voltage on the suppressor grid causes redistribution of the anode and screen grid currents.

The following example will illustrate the operation of the transitron circuit. Assume that oscillations with an amplitude of 10 v have commenced in the tuned circuit and that at the given moment the upper plate of capacitor C is charged negatively. D. c. voltages of anode, screen grid and suppressor grid are, respectively, given by the following values: 100 v, 200 v and 0 v. On a negative half-wave, the voltage of 10 volts will decrease the screen-grid voltage from 200 down to 190 v and, as a result, the screen-grid current will be slightly reduced. But at the same moment, a simultaneous change of suppressor-grid voltage will take place, this voltage no longer having a zero value but increasing in the negative direction to —10 v. This will sharply reduce the anode current, the reduction of this current, in its turn, causing a considerable increase of the screen-grid current. Thus, a reduction of the screen-grid voltage (within certain limits) will cause an increase of screen-grid current, and vice versa. In the circuit being examined, it may be seen that when the upper plate of capacitor C is charged negatively, an increase of current I_{g2} will replenish this charge, thus compensating for the energy loss in the tuned circuit and sustaining self-oscillation.

The transitron oscillator is capable of generating a. c. power in a frequency range extending from several dozens of cps to many megacycles. The control grid does not participate in the oscillatory process but is supplied with a certain value of negative bias voltage in order to set the correct operating condition of the oscillator.

A valve oscillator, when coupled to an aerial, constitutes a valve radio transmitter. The transmitter may be directly connected to its aerial, and then the aerial will act as a part of the tuned circuit of the oscillator. Alternatively, the transmitter may be coupled to the aerial by means of any one of the known types of coupling arrangements (inductive, capacitive or autotransformer coupling), and then the tuned aerial circuit will become an independent circuit, coupled to the tuned anode circuit of the oscillator.

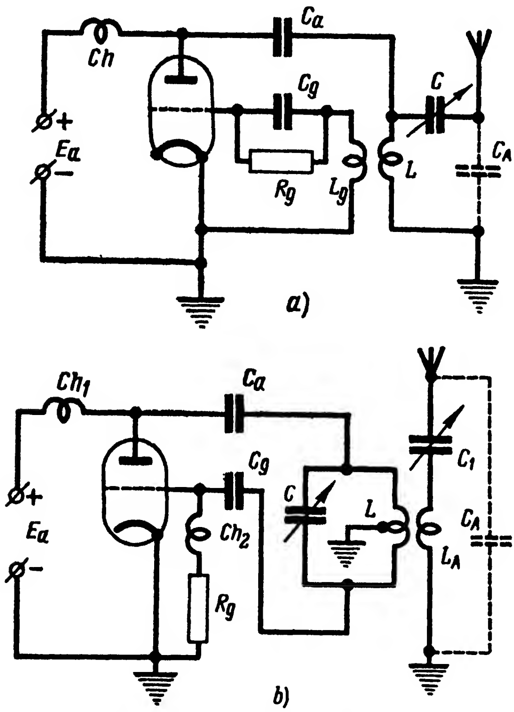


Fig. 180. Self-excited transmitters employing: *a)* a simple aerial circuit; *b)* a complex aerial circuit

A transmitter with a directly-connected aerial is said to employ a *simple aerial circuit.* On the other hand, a transmitter coupled to its aerial is said to employ a *complex aerial circuit.* In most cases, such transmitters are inductively-coupled to their aerials.

Radio transmitters employ various types of valve oscillators. Fig. 180*a* shows a transmitter based on an oscillator with inductive feedback, employing a simple aerial circuit and parallel anode feed. In this transmitter, the capacitance of the tuned circuit is comprised of capacitance C and the capacitance C_A of the aerial. Fig. 180*b* shows a transmitter with autotransformer feedback and a complex aerial circuit. The oscillator is inductively coupled to the aerial. Capacitor C_1 serves to tune the aerial circuit in resonance with the oscillator frequency.

A slightly greater aerial power P_A obtained with the simple aerial circuit is the advantage of this circuit. This circuit, however, is noted for the following serious disadvantage. Under class B or class C operating conditions of the oscillator, when the anode current pulses are not sinusoidal, the directly-connected aerial

radiates numerous harmonics. The radiation of these harmonics creates interference to other stations operating on frequencies which are by 2, 3, 4, etc., times higher than the fundamental frequency of the given transmitter. For instance, if the transmitter operates on a fundamental frequency of 3,000 kc, which corresponds to a wavelength of 100 metres, the simple aerial circuit will also give a considerable radiation of the second harmonic ($f = 6{,}000$ kc or $\lambda = 50$ m), the third harmonic (9,000 kc or $\lambda = 33.3$ m), etc.

The harmonic radiation is weaker than the radiation of the fundamental frequency, but, still, it is sufficient to interfere with the reception of other stations operating on waves whose frequencies are close to the frequencies of these harmonics.

In transmitters employing complex aerial circuits, the tuned aerial circuit is a secondary circuit, tuned to resonance with the fundamental frequency of the transmitter. Because of this, the harmonics in such a circuit are considerably attenuated. It is customary to say, that the complex aerial circuit *filters out harmonics.* The disadvantage of the complex circuit is the reduction of the aerial power it introduces. Practically, P_A is equal from 0.5 to 0.9 P_k, where P_k is the power developed in the tuned circuit of the oscillator. The looser the coupling between the transmitter and the aerial, the lower the value of P_A, but, then, the harmonic radiation is also lower. The coupling between the transmitter and the aerial is made variable as a rule. The aerial coil which couples the transmitter to the aerial usually comprises a small number of turns.

Single-valve self-excited transmitters have the simplest circuit and construction. They are noted, however, for the following serious disadvantage. The frequency of oscillations of a self-excited oscillator is determined by the natural frequency of the tuned circuit of the oscillator. In transmitters based on self-excited oscillators, the aerial is either a part of the tuned anode circuit, or else is closely coupled to this circuit. Hence, the transmitter frequency fluctuates when the aerial parameters are varied for one reason or another, for instance — for the reason of aerial swinging by the wind, which changes the aerial capacitance. This dependence of frequency upon aerial parameters is lessened in transmitters employing complex aerial circuits, if the coupling between the transmitter and the aerial is made loose, but then a loose coupling decreases the transmitter power output to the aerial.

As follows from the above discussion, when a practical transmitter consists of a single-stage self-excited oscillator, its operating frequency will not be stable. Under certain conditions, the reception of signals of such a transmitter becomes quite impossible. As a matter of fact, the tuning scale of this type of transmitter cannot be exactly calibrated in wavelength or frequency. Because of this, such a transmitter cannot be tuned to an exact

required frequency and will create interference to the reception of other stations operating on frequencies which are close to the expected frequency of the offending transmitter.

This, of course, is intolerable, particularly in our days, when *frequency stability is the main requirement that a radio transmitter has to meet.*

This requirement is not met by a transmitter using a single self-excited stage and generally referred to as *oscillator-transmitter.* There was a time when oscillator-transmitters were the only types of transmitters in existence. But that, however, was long ago, when radio engineering standards were not strict and when there were but few radio stations on the air. But this era is now over; and, at present, oscillator-transmitters are allowed to go on the air only on ultra-short-wave bands (metric and still shorter waves), where the inherent instability of such transmitters is tolerated to a certain extent.

86. M.O.P.A. TRANSMITTERS

The frequency of waves radiated by modern transmitters is very stable, because the output stages of such transmitters are separately excited.

A separately-excited transmitter consists of an exciter, known as a master oscillator and represented by a self-excited oscillator, and of one or more power amplifier stages. Radio engineers call this type of transmitters M.O.P.A. transmitters, which is an abbreviation of "Master Oscillator-Power Amplifier" transmitters.

In a transmitter of this type, the master oscillator can utilise any one of the oscillator circuits already studied. The power amplifier is a high-frequency amplifier stage terminating with a tuned circuit, the latter serving as the load of the amplifier. Fig. 181 shows the simplest version of a M.O.P.A. transmitter, employing only one stage of amplification. Here, the master oscillator employs a triode with autotransformer feedback and parallel anode feed. The power amplifier consists of a tetrode with parallel anode feed. The power amplifier is coupled to the master oscillator by means of autotransformer coupling, and is inductively coupled to the aerial. We shall refer to this circuit when studying the peculiarities of a M.O.P.A. transmitter.

The function of the master oscillator is to generate oscillations of a required frequency and of the highest order of stability.

These oscillations are fed to the next stage for the purpose of amplification. The main function of the power amplifier is that of isolating the master oscillator from the aerial, so that changes of aerial parameters would not affect the frequency generated by the master oscillator. Besides this, the power amplifier also serves to amplify the power of oscillations fed to its grid by the master oscillator.

The power amplifier stage usually employs a tetrode, or a pentode. A triode is undesirable in this stage, because the parasitic capacitance C_{ag}, existing between the anode and the grid of a triode, accounts for the following deleterious effects:

1) the aerial capacity, acting through the C_{ag} capacitance, exerts an undesirable influence upon the tuned circuit of the master oscillator, thus causing frequency instability;

2) the oscillations generated by the constantly operating master oscillator will pass on to the aerial through the same capacitance

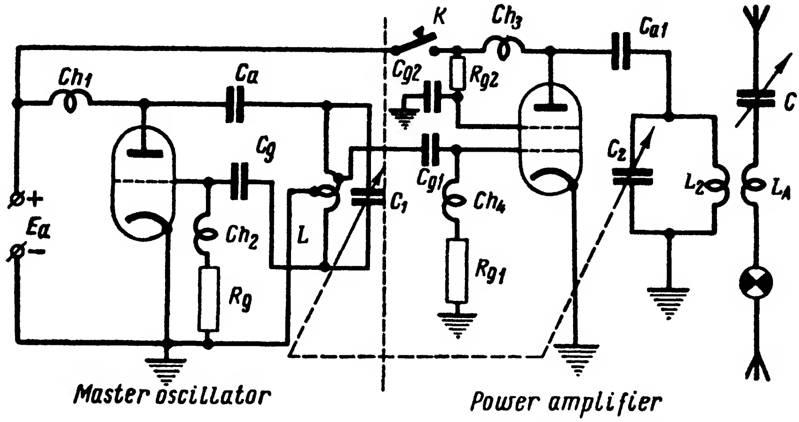


Fig. 181. Separately-excited transmitter employing a complex aerial circuit

C_{ag} and will be radiated by the aerial even when the sending telegraph-key, K, is open. This would give rise to the transmission of the so-called "negative signals", occurring during the intervals between dots and dashes of the telegraph transmission;

3) interelectrode capacitance C_{ag} provides a very undesirable capacitive coupling between the anode and grid of the power amplifier stage. As a result, this stage is converted into a self-excited tuned anode-tuned grid oscillator. The frequency of oscillations generated by the power amplifier in this case will be determined by the parameters of tuned circuit L_2C_2, which is coupled to the aerial and, because of this, has poor frequency stability. In other words, the power amplifier will be turned into a self-excited oscillator with inductive coupling to the aerial, and we already know all the disadvantages of such a circuit (see Sec. 85).

Despite all this, power amplifier stages sometimes employ triode valves. However, in this case the triodes are *neutralised* by incorporating special neutralising capacitors in the amplifier stage. Proper adjustment of such a capacitor neutralises the C_{ag} capacitance and eliminates its deleterious effect.

Still, tetrodes and pentodes are more frequently used in power amplifier stages. Proper arrangement of wiring of the stages eliminates the external capacitance between anode and grid circuits. This prevents the undesirable effects observed with triodes and makes neutralisation unnecessary.

Master oscillator stages are usually single-ended; the push-pull arrangement is seldom resorted to and only when the oscillator operates on very short waves. The tuned circuit of the master oscillator is coupled to the grid circuit of the power amplifier by means of either inductive or autotransformer coupling, although capacitive coupling is also used at times.

Requirements of frequency stability call for the loosest coupling, consistent with sufficient energy transfer from the master-oscillator anode-circuit to the power-amplifier grid. In this case, loose coupling provides a better degree of isolation between the power amplifier and the master oscillator, and, hence, decreases the undesirable influence of the first upon the second. However, the looser the coupling, the greater must be the power of the master oscillator to properly drive the power amplifier. In practical cases, the power of the master oscillator amounts to 10-50% of the power of the output stage. This final stage can be either of single-ended or push-pull variety. A push-pull power amplifier stage gives a considerable increase of power and is particularly desirable on very short waves. The power output of the final stage is sometimes increased by employing several parallel-connected valves in the stage, but this also increases the interelectrode capacitances, which is particularly undesirable in short-wave transmitters.

For the purpose of reduction of harmonic radiation, power amplifiers employ complex aerial circuits. Then the tuned anode circuit of the amplifier becomes known as the intermediate tuned circuit. For the purpose of tuning convenience, variable capacitors C_1 and C_2 of master oscillator and amplifier are sometimes operated by a common shaft (ganged).

The rotors of the two variable capacitors are, thus, rotated as a unit and the two circuits are tuned simultaneously. This, however, is resorted to only in low-power transmitters, where the coupling of the intermediate tuned circuit with the aerial is fixed. Higher-power transmitters use variable aerial coupling. The aerial circuit of a radio transmitter is tuned by a variometer or by a variable capacitor. The role of the aerial tuning-indicator is performed by a small incandescent lamp, or by a neon lamp, in many low-power transmitters. Higher power transmitters employ hot-wire or thermocouple ammeters for the purpose, the ammeters giving direct readings of the aerial current. The aerial tuning capacitor is always provided with a separate handle. This capacitor cannot be ganged with the master oscillator and power amplifier tuned-circuit capacitors, because the parameters of an aerial are not constant.

A transmitter is tuned in the following way. If the tuning capacitors of the master oscillator and power amplifier are ganged and the aerial coupling is fixed, first the ganged capacitors are turned and their tuning scale is set to a required wave, after which the aerial control is adjusted for maximum aerial current. When dealing with transmitters of higher power, provided with a separate adjustment of the intermediate tuned circuit and with a variable aerial coupling, the transmitter is first set to a required wave by adjusting the master oscillator tuned circuit, whereupon the amplifier tuned circuit is tuned to resonance.

The condition of resonance is shown by a corresponding reading of some meter which must be introduced in the amplifier circuit, the latter being kept loosely coupled to the aerial. Once the intermediate circuit has been tuned to resonance, the aerial coupling is tightened, after which the aerial circuit is tuned. Maximum aerial power is obtained by setting the aerial coupling to its optimum value and by trimming the intermediate tuned circuit and aerial tuning. This trimming becomes necessary, because the change of coupling and the change of parameters of one of the tuned circuits affect the parameters of the other tuned circuit and upset resonance. During this tuning procedure the anode of the amplifier valve must be watched; too tight a coupling of the aerial is likely to cause overheating of the anode.

87. ELECTRON-COUPLED OSCILLATORS

Electron-coupled oscillators are widely employed by radio transmitters of low and medium power, as well as by heterodyne circuits used in radio receivers and measuring apparatus. An electron-coupled oscillator is a clever application of a single valve as a master oscillator and a power amplifier. Such a combination represents, in fact, a one-valve M.O.P.A. transmitter of a frequency stability higher than the stability of a simple self-excited oscillator. Fig. 182*a* shows the circuit of an electron-coupled oscillator employing a filamentary valve. The circuit can use a tetrode, a pentode or a more complex valve. The filament, the control grid and the screen grid of the valve act as a triode oscillator, the screen grid functioning as the anode in such triode. This master oscillator usually employs autotransformer feedback and series anode-feed (cathode-coupled circuit shown in Fig. 175*c*). This circuit arrangement is reproduced in Fig. 182*a*, although other circuit arrangements are possible. The anode end of the tuned circuit L_1C_1, referred to as the ***inner tuned circuit***, is connected to the negative terminal of the anode power supply. The filament current flows through a part of the tuned circuit coil. High-frequency choke Ch_1, connected into the positive filament wire, prevents short-circuiting this part of the

coil (in respect to the alternating anode current) by the filament battery. Dropping resistor R_{g2} lowers the value of voltage fed to the screen grid of the valve, while capacitor C_{g2} by-passes the a.c. component of anode (i.e., of screen grid) current of the master oscillator.

Oscillations set up in the tuned-circuit L_1C_1 of the master oscillator are amplified by the whole valve, which is included in the power

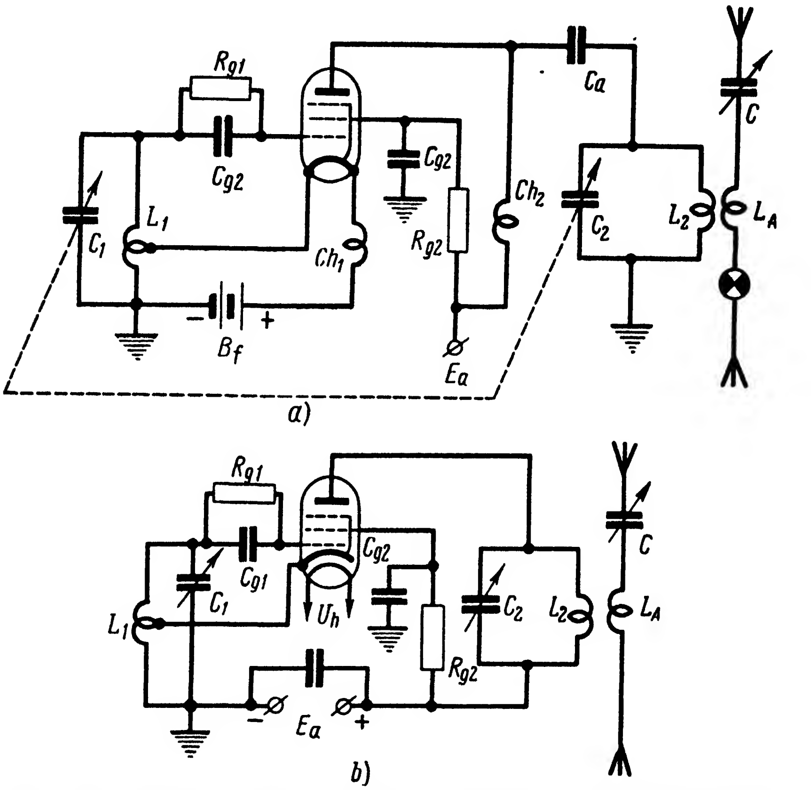


Fig. 182. Electron-coupled oscillator employing: *a)* a filamentary valve: *b)* a cathode-type valve

amplifier circuit and employs parallel anode feed (although it can also use series anode feed, as an alternative). Amplified oscillations are obtained in the anode tuned circuit L_2C_2, referred to as the *outer tuned circuit* and coupled to the aerial. The circuit arrangement of an electron-coupled oscillator is somewhat modified when a cathode-type valve is used. This is logical, because the heater circuit is independent of other circuits in such a valve, which makes the high-frequency choke unnecessary (Fig. 182*b*). However, in short-wave transmitters high-frequency chokes are connected,

as a rule, in both heater wires. This is resorted to because one side of the heater supply is usually connected to the common negative terminal and has a considerable capacitance in respect to the chassis. Besides, quite a large parasitic capacitance also exists between the cathode and the heater. This capacitance shunts a part of coil L_1, when the chokes are omitted. It is customary to by-pass high-frequency currents from the cathode to the heater by means of a capacitor having a value of several hundred or thousand picofarads. If this is not done, the insulation between the cathode and the heater may be punctured by the considerable high-frequency voltage developed across this part of the circuit. Tuning capacitors C_1 and C_2 are usually ganged.

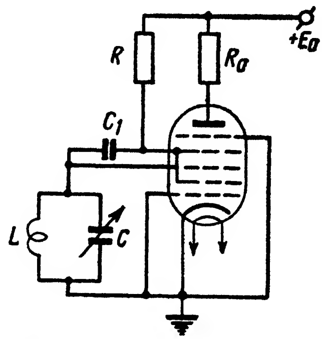


Fig. 183. Electron-coupled transitron oscillator

The operation of an electron-coupled oscillator, as explained above, is very simplified. It must be considered that the a.c. component of the anode current of the power amplifier (i.e., the a.c. component of the main current) flows through tuned circuit L_1C_1 thus facilitating self-excitation of the master oscillator. If the circuit of the main anode is opened, the power of oscillations in tuned circuit L_1C_1 is lowered and, in some cases, the master oscillator will stop functioning. Thus, in an electron-coupled oscillator, power amplifier oscillations reach the master oscillator and help to keep it in the state of oscillation.

In the usual two-valve M.O.P.A. circuit, the a. c. component of the power amplifier anode current does not reach the master oscillator, which distinguishes such a circuit from an electron-coupled oscillator. It should not be thought, however, that parasitic capacitive coupling exists between tuned circuits L_1C_1 and L_2C_2 in an electron-coupled oscillator; these circuits are coupled to each other through the common electron stream only, which accounts for the name of "electron-coupled oscillator". Because of the absence of parasitic coupling, changes of parameters of tuned circuit L_2C_2 exert but a very slight influence upon the parameters of the inner tuned circuit, which accounts for high frequency stability of the master-oscillator section in an electron-coupled oscillator stage.

Low-power electron-coupled local oscillators (heterodynes), employed in radio receivers, sometimes employ an ohmic resistor or a high-frequency choke in place of the tuned circuit L_2C_2. Certain electron-coupled heterodynes employ transitron exciters and are based on heptode valves, as a rule. A circuit of such transitron heterodyne, using a type 6A7 valve, is shown in Fig. 183. The operation of this circuit is similar to that of the circuit employing

a pentode shown in Fig. 179*b*. The roles of the screen grid and suppressor grid of the pentode are performed by the second and the third grids of the valve, the fourth grid acting as the exciter-valve anode. The alternating voltage developed across the tuned circuit is amplified by the whole valve, and the amplified voltage is taken off from across anode load resistor R_a.

88. FREQUENCY CONTROL

The importance of frequency stability in modern radio transmitters has already been discussed above. The simplest method of obtaining frequency stability is that of employing separate excitation, and complex aerial coupling circuits. This, however, is not sufficient, because the exciter stage (the master oscillator) is apt to change its frequency owing to various capacitive and thermal effects and also owing to the instability of the power supply voltages. Hence, it becomes necessary to stabilise, i.e., to control, the frequency of the master oscillator. Two general methods of frequency control are used: *the parametric frequency control* (requiring no crystal) and *crystal frequency control.*

The parametric frequency control is based on the weakening of various external factors influencing the oscillator frequency. It is also based on the selection of appropriate parameters and circuit elements securing minimum frequency changes. One of the parametric frequency control measures consists in operating the oscillator stage at the longest possible wave (not shorter than 40 metres), in using a high value of capacitance in the tuned circuit, and in the subsequent *frequency multiplication.*

It is a fact that the frequency stability of a valve oscillator generally becomes poorer on shorter waves. Suppose that a slight change of the tuned circuit capacitance or of the operating condition of a valve oscillator changes the frequency by 1%. If the oscillator frequency were originally 200 kc (wavelength = 1,500 metres), the resultant frequency change will be only 2 kc, which is a comparatively small value. A radio receiver tuned to such a transmitter will still reproduce the signal, despite the 2-kc detuning. However, if the transmitter were operating on 5,000 kc (wavelength = 60 metres), the 1% frequency change would be equivalent to a 50-kc change and the signal would no longer be audible in a radio receiver tuned to the original transmitter frequency of 5,000 kc.

The advantage of employing a high value of capacitance in a tuned circuit can be readily shown by the following example. Assume that, for one reason or another, the capacitance of the tuned circuit changes by 0.5 pf, causing a corresponding frequency change. If the tuned circuit capacitance were originally equal to 50 pf, such a change becomes equivalent to 1%. However, if the original

value of capacitance were 500 pf, a 0.5-pf change will be equal only to 0.1%, which will affect the frequency to a much lesser degree. It is, however, undesirable to employ tuned circuits with too high a value of capacitance, because the inductance employed by such circuits has to be made small, which reduces the power output of the oscillator; a good practical compromise is had when the value of capacitance, expressed in picofarads, is 5 times as large as the wavelength, expressed in metres.

Frequency multiplication becomes necessary when the master oscillator of a short-wave transmitter operates on a longer wave than that radiated by the station. The frequency of the master oscillator may be doubled with the help of a single-ended circuit. A frequency-doubler stage resembles an ordinary amplifier stage but operates under different condition. The bias voltage of a doubler stage is 2-3 times as high as that of an amplifier stage, and the exciting voltage applied to the grid of the doubler is respectively higher. Under such operating condition, the second harmonic is most pronounced in anode current pulses. It is to this second harmonic that the tuned anode circuit of such a doubler stage is tuned.

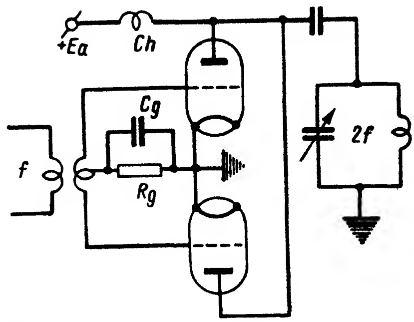


Fig. 184. "Push-push" frequency doubler

The circuit thus adjusted presents high impedance only to the second harmonic (a case of parallel resonance). Consequently, high-power oscillations are set up in such a tuned circuit, the frequency of these oscillations being twice as high as the frequency of the oscillations applied to the grid circuit of the stage. The tuned circuit adjusted to the second harmonic will present but a low value of impedance to the fundamental frequency, and no amplification of this frequency will take place in the doubler. The useful power output of a doubler stage is low and cannot exceed 50% of the power output this stage would give if it were operated as a power amplifier.

Another type of frequency doubler is shown in Fig. 184. The operation of this circuit is based on the fact that only even harmonics are present in the common anode circuit of a push-pull stage. The grid circuit of such a doubler is similar to that of a push-pull amplifier and, hence, the two valves used by this doubler operate with a 180° phase shift (i.e., first one valve, then the other). The anodes of the two valves are connected in parallel and a tuned circuit, adjusted to the second harmonic, is connected into the common anode wire. Doublers of this type are sometimes called "push-push" doublers.

In a push-push doubler, the bias voltage and grid excitation are ***so selected that the second harmonic,*** **appearing in the tuned anode circuit, is most pronounced.**

Apart from frequency doublers, frequency triplers are also used. The anode circuit of a frequency tripler is tuned to the third harmonic of the fundamental frequency applied to the tripler grid. Such operating condition considerably reduces the output power of the stage. The usual electron-coupled oscillator can be used as a frequency doubler or tripler by the simple expedient of tuning its outer tuned circuit either to the second or the third harmonic of its inner circuit.

The use of several frequency doublers or triplers in a radio transmitter offers a multi-fold frequency multiplication, the output stage of such a transmitter developing very short, even ultra-short, waves with good frequency stability, although the master oscillator of the transmitter works on a much longer wave.

Changes of supply voltage exert a profound influence upon the frequency stability of an oscillator. If the oscillator operates from a.c. mains, a voltage stabilising circuit should be incorporated in the rectifier. It is best to feed such an oscillator from a separate rectifier, in any case. For the sake of better stability, oscillators are made to operate with reduced anode voltage. In this case, the voltage is lowered by connecting a dropping resistor into the anode circuit of an oscillator, the value of such resistor ranging from several thousand ohms to several tens of thousands of ohms. An increase of the grid-resistor value in such an oscillator also improves stability. This measure, of course, results in lowering the output power of the oscillator, just as it does when the anode voltage is decreased.

Thermal effects, affecting the frequency of oscillators, are most difficult to counteract. When an oscillator valve is put into operation, gradual heating of its electrodes, as well as the heating of various external circuit components associated with the given valve, slowly changes the parameters of the tuned circuit (for instance — increase its capacitance). This brings about what is known as frequency drift, i.e., a slow creeping of the oscillator frequency, such creeping being observed for a considerable length of time — in some cases for half an hour or so after the oscillator had been switched on. This is why it is advisable to switch on the oscillator a long time before the transmitter goes on the air, and to keep it switched on during pauses in transmission. In higher-power transmitters, the master oscillator stage is enclosed in a thermostat. A thermostat is an enclosed box provided with thermal-insulating walls, electric heaters and automatic relays, the electrical circuits so arranged to operate that a constant temperature is maintained within the box. Tuned-circuit components (capacitor and coil) are temperature-compensated, i.e., they are made to resist such capacitance and inductance changes that the natural heating could produce. The simplest

method of temperature compensation is that of including an auxiliary *tycond capacitor* into a tuned circuit (see Sec. 14).

The components of an operating oscillator should be secured against mechanical vibration. If the tuned circuit coil, capacitor or associated wires are not securely fixed, even slight vibration can cause considerable frequency changes. This is why oscillator stages, particularly, should have rigid construction and wiring, especially in portable and mobile radio stations. An oscillator should be loosely coupled to the following stage, otherwise this stage will be likely to affect the oscillator frequency. It is very desirable that such a stage, driven by the oscillator, operates without grid current. In high-power transmitters stages, called the *buffer stages*, are always used to isolate the oscillator from the influence of the amplifier-stages developing high levels of power. In some cases, when it is required to eliminate various capacitive effects upon the oscillator, the whole oscillator-stage is shielded.

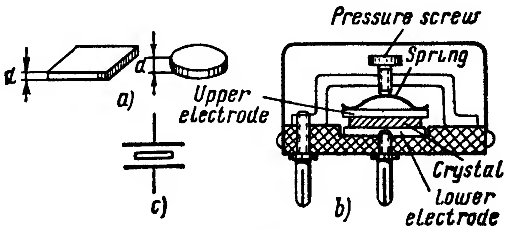


Fig. 185. Quartz plate and crystal holder

Crystal-controlled oscillators provide a much higher stability than oscillators of other types. Such oscillators are quite simple in design and require the inclusion of but one auxiliary component—a quartz plate, usually called the crystal, into their circuits. In comparison with the parametric frequency control, the crystal control is noted for the following disadvantage: a crystal, as a rule, controls only one frequency, while the parametric control stabilises all the frequencies of the range covered by an oscillator.*

Another disadvantage of crystal control is its comparatively high cost and the fragility of the crystal, which can be easily fractured if carelessly handled.

Crystal plates used by oscillators are cut out of quartz crystals (rock crystals) or quartz pebbles and have rectangular or round shape (Fig. 185*a*). The crystal plate is then placed in a crystal-holder (Fig. 185*b*) between two metal plates. An alternative way is that of plating the quartz surfaces with metal layers. This forms a capacitor in which the quartz plate serves as a dielectric. The schematic representation of a crystal is shown in Fig. 185*c*.

Quartz crystals possess piezoelectric properties, discussed in Chapter VI. This means that when alternating voltage is applied

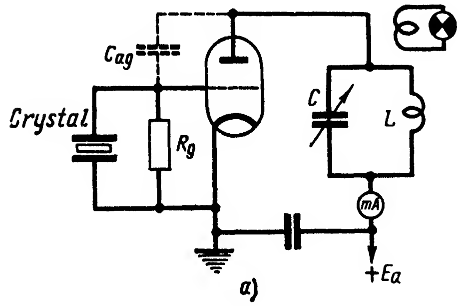
* It is possible to devise crystal-control circuits exerting control over a continuously tuned frequency range, but such circuits are comparatively complicated.

to a quartz crystal, the latter begins mechanical oscillations, i.e., it will periodically contract and expand.

A quartz plate has several natural frequencies, depending upon the dimensions of the plate. Frequency control usually utilises "the thickness oscillations" of the plate, the frequency of such oscillations depending upon the plate thickness. The respective wavelength of an electromagnetic wave, measured in metres, is determined by plate thickness d, given in millimetres in accordance with the following formula:

$$\lambda = 120\,d.$$

For instance, when $d = 0.5$ mm, the wavelength is given by the following: $\lambda = 120 \times 0.5 = 60$ m. Although the constant is given as 120 in the given case, it varies from 100 to 140 for various types of quartz plates.



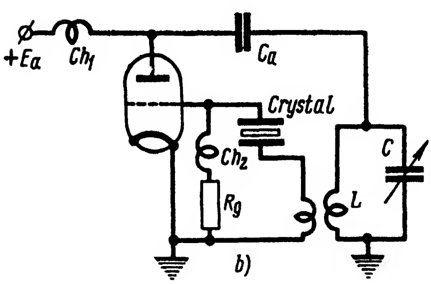


Fig. 186. Crystal oscillator circuits, one employing a feedback through anode grid capacitance *(a)*, and the other employing an additional inductive feedback *(b)*

The dimensions and properties of quartz vary with temperature, which changes the natural frequency of a quartz plate. Formerly, this effect was counteracted by placing crystal plates into thermostats. Modern engineering practice prefers to counteract this effect by cutting crystal plates in a special way (the so-called zero-angle cut). In a zero-angle-cut crystal, the oscillation frequency is almost totally independent of temperature.

The strongest undamped oscillations of a crystal plate are obtained when the frequency of external alternating e.m.f., applied to the crystal, is equal to the natural frequency of the crystal, i.e., when a state of resonance sets in. A crystal has a very sharp resonance and, hence, very low damping.

The amplitude of crystal plate oscillations becomes quite negligible with the slightest difference between the frequency of external e.m.f. and the natural frequency of the plate. It is this property of crystal that is utilised for frequency control. A quartz plate is equivalent to an oscillatory circuit possessing very low damping and very stable frequency.

One of the most popular crystal-controlled oscillators, in which the crystal is connected between the grid and cathode of oscillator valve, is shown in Fig. 186*a*. In effect, this is a tuned-anode tuned-grid circuit, in which the crystal takes the place of the tuned-grid part of the circuit. In this circuit, the feedback path is the anode-grid capacitance C_{ag}. The tuned-anode circuit is adjusted to the frequency of the crystal. The frequency of voltage, fed by the crystal to the grid of the oscillator valve, is very stable thus making for the high stability of the a.c. component of the anode current and for a similarly high stability of oscillations in the tuned-anode circuit. When the latter circuit is slightly detuned, the oscillations are still sustained. However, if this circuit is detuned further, its impedance will drop, the amplification will be reduced, and the oscillations fed to the grid circuit from the anode via feedback will not be sufficiently strong to maintain the crystal in the state of continuous oscillations.

Thus, an oscillator employing a crystal in its grid circuit is capable of generating high frequency oscillations only over a narrow sector of anode circuit tuning range. The presence of oscillations is detected by an indicator, such as a small electric lamp inductively coupled to the tuned circuit.

The presence of oscillations is also indicated by the reduction of anode current, as read by a milliammeter connected in series with the anode supply source. The latter effect is attributed to the following. When oscillations are present, alternating voltage is developed in the grid circuit and causes a grid current flow. This current, flowing through resistor R_g, builds up negative bias voltage on the grid, thus reducing the direct anode current. In this case, no capacitor is required in the grid circuit, because direct grid current cannot pass through the crystal. An oscillator employing a crystal in its grid-circuit can use either parallel or series anode feed.

Some quartz crystals are not very active and oscillate with difficulty. Moreover, in oscillators using crystals in their grid circuits the value of feedback becomes insufficient on longer waves. In such cases, the operation of a crystal-controlled oscillator is facilitated by incorporating additional feedback into the circuit, such feedback simultaneously increasing the output power of the stage.

The value of this additional feedback is so adjusted that the circuit does not oscillate without the crystal, i.e., does not oscillate on frequencies different from the crystal frequency, but, at the same time, is close to the oscillatory condition. When the circuit thus adjusted is tuned to the crystal frequency and the crystal begins to function, applying additional excitation to the grid, the circuit will oscillate. A circuit of this type, incorporating the additional feedback of inductive variety, is shown in Fig. 186*b*.

On waves shorter than 30 metres, a quartz plate has to be very thin. The plate becomes extremely fragile and is easily destroyed

by overloads. The maximum permissible power of a crystal-controlled oscillator depends upon the thickness and the area of the quartz plate it employs. This power — i.e., the power in the tuned circuit of the oscillator — should not exceed 1-2 watts per 1 sq cm of plate area for crystals whose thickness is equal to 0.5 mm. Short waves in the range of 20-10 metres, as well as ultra-short waves, have to be stabilised by incorporating the crystal control in a low-frequency oscillator and by resorting to frequency multiplication in the following transmitter stages.

Tourmaline plates, used in place of quartz plates, make it possible to control directly even ultra-high frequency oscillators without frequency multiplication. Besides this, tourmaline plates can withstand higher power than quartz plates. Tourmaline plates, however, are more expensive than quartz plates and, because of this, are only seldom used.

89. TELEGRAPH KEYING OF RADIO TRANSMITTERS

By telegraph keying of a radio transmitter is meant the process of starting and stopping the radiation of the transmitter by means of a telegraph key, thus making the transmitter send dots and dashes of a radio telegraph message.

Speaking generally, there are two methods of keying. In the first of these methods, the transmitter stops radiating altogether when its telegraph key is opened. In the second method, opening the key prevents the transmitter from radiation on an assigned operating wave but the radiation continues on another wave, called the negative wave. Thus, in the latter of the two methods the telegraph dots and dashes are sent on the operating wave, the transmitter radiating continuously on another wave during the intervals between such dots and dashes.

A telegraph key can be connected to various points of a transmitter circuit. It is a poor practice, however, to key the master oscillator, because this would cause instability of its frequency. Hence, it is generally preferable to connect the key to some point of the circuit of one of the amplifiers. Either the final or an intermediate amplifier may be successfully keyed. If the transmitter is of low-power variety and operates on comparatively low anode voltage, the simplest way of keying is that of breaking and making the anode circuit of one of the valves by means of a telegraph key. Alternatively, the key may be arranged to close and open the screen-grid circuit of an amplifier valve. However, before connecting the key either into the anode or screen-grid circuit, the question of operator's safety must be considered. If a potential of several hundred volts exists at the key, operator's life will be endangered, even if the key is provided with protective cover. Apart from this consideration, keying a high-voltage lead is

also undesirable for the following technical reason. When a key is arranged to break a high-voltage circuit, excessive sparking occurs at the contacts of the key. This sparking burns the key contacts and; besides this, creates interference to radio reception in the neighbourhood. Such sparking can be largely suppressed by connecting special capacitors and resistors across key contacts.

One popular way is that of connecting the key into the control grid circuit of an amplifier valve. Care should be taken, however, that the key is not connected into a wire carrying high frequency

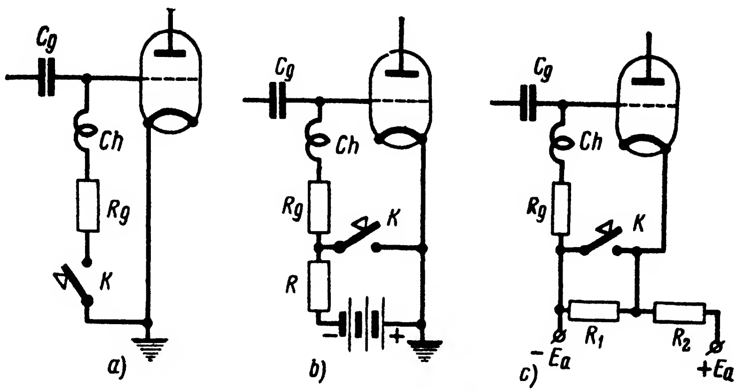


Fig. 187. Various methods of grid-circuit keying

currents, as this is apt to bring in the so-called "hand capacitance" effect, when the proximity of the operator's hand to a high-frequency circuit affects the tuning of the circuit.

One way of grid-circuit keying is shown in Fig. 187*a*, where the key is placed into the lead carrying the d.c. component of the grid current only. In this arrangement, opening of the key results in accumulation of electrons on the grid, which cuts off the anode current of the valve. However, this method is not applicable to valves with "left-handed" characteristics, where the negative potential formed on the grid by electron accumulation is insufficient to cause anode current cutoff.

A more reliable way of grid-circuit keying is shown in Fig. 187*b* and 187*c*. In the two alternative circuits given here, opening of the key automatically applies to the grid of the valve a sufficiently high negative potential to cut off the anode current. In the first case (Fig. 187*b*), the negative potential is supplied by a separate source of voltage. In the second case (Fig. 187*c*), the required cutoff potential is built up across a part of a voltage divider.

When the telegraph key is located at a considerable distance from the radio transmitter (which is generally so in the case of high-power transmitters), the transmitter radiation is controlled by means of a

special relay. Such a relay is installed in the transmitter and its contacts make and break some part of the transmitter circuit, causing the latter to send dots and dashes. The relay, in its turn, is controlled by a remotely-located key, worked manually by a radio operator, or else worked automatically.

90. TRANSMITTING VALVES

Low-power oscillators and transmitters can employ any types of receiving triodes, tetrodes and pentodes, which we have already discussed under Chapter IV. Among the small-dimensional glass valves, valves specially manufactured for application in low-power transmitters were represented by type СБ-245 tetrode and type СО-257 pentode. These two valves differ from the usual receiving valves in that the terminal brought out at the top of their envelope is the anode and not the control grid.

Medium-power and high-power transmitters employ special transmitting valves. The filaments of such valves are made of pure tungsten. The only exception are the transmitting valves designed for operation on anode voltage lower than 1,500 volts; these valves employ oxide-coated or carbidised cathodes.

The anodes of medium-power transmitting valves are often made of nickel and are usually blackened so that their heat loss is improved. Valves designed for higher power ratings employ molybdenum and tantalum anodes. In highest-power water-cooled valves the anodes are made of copper or chromium steel. The grids are made of molybdenum. Medium-power valves not intended for operation on ultra-short waves are provided with pin-type bases. In such a valve, the anode is usually brought out through the top of the glass envelope. In higher-power valves, especially in those designed for ultra-short wave service, electrode terminals are often brought out through various parts of the glass envelope and have the shape of special contacts or flexible leads terminating with lugs. Such valves are manufactured either totally without a base, or else are provided with a base which carries only the filament, screen-grid and suppressor-grid pins, the anode and control grid terminals being brought out through other parts of the valve.

The marking formula of a transmitting valve begins with Russian letter Г. In the old-type marking system, this letter was followed either by the number of the plant job lot, or else by a letter indicating the wavelength range for which the valve was designed, letter Д standing for waves longer than 200 metres, letter К — for short waves of 15 metres and longer, and letter У — for ultra-short waves. In case of tetrodes, a third letter Э was included in the marking formula. The number standing after the letters indicated the useful power output which the valve was capable of giving under normal operating condi-

tion in the wavelength range for which it was designed. Thus, for example, a marking formula ГК-20, attached to a certain valve, indicated that the valve was designed for transmitting-duty on short waves and was capable of giving an output of 20 watts.

The new system of marking is somewhat different. Here, the marking formula either begins with letters ГК, meaning that a given transmitting valve is designed to operate on long and short waves (frequencies below 25 mc), or with letters ГУ, meaning that the valve is designed to operate on ultra-short waves (frequencies between 25 and 600 mc). Corresponding designation for valves designed for operation on centimetric waves (frequencies above 600 mc) is ГС, for valves designed for modulator service — ГМ, and for valves intended for pulse operation — ГИ. In the new marking formula, the letters given above are followed by a number to distinguish various types of valves from each other. The new marking system notwithstanding, many valves used at present are still manufactured with old-type marking, which was explained above.

It should be noted that valves designed for short-wave operation are capable of giving good performance on long waves. Certain long-wave valves will give a good account of themselves on short waves, providing their output power is lowered. Thus, type ГК valves can be sometimes used in the ultra-short wave range.

Transmitting triodes designed for medium and short waves are noted for high values of amplification factor μ (fine grid) and for "right-handed" characteristics. This allows to drive such valves to full output even with a low value of excitation voltage. The same feature facilitates setting them into oscillation when the valves are used as self-excited oscillators, but in this case the operation is accompanied by large values of grid current.

Transmitting tetrodes are chiefly employed as power amplifier valves and also as electron-coupled oscillators. The dynatron effect in these valves is eliminated by keeping the screen-grid voltage at a low value (15-20% of the anode voltage). Transmitting pentodes and beam tetrodes have found a particularly wide application and, because of their superior qualities, are rapidly replacing triodes and tetrodes.

The suppressor grid always has an independent terminal in all transmitting pentodes, because in radio telegraph transmitters this grid is sometimes made positive to increase the power output of the stage, while in radio telephone transmitters modulation is occasionally applied to the suppressor grid.

Below are listed the most popular transmitting valves, designed for power output up to 1,000 watts.

For long and short waves: triode ГК-20, tetrodes ГКЭ-100 and ГКЭ-500, beam tetrode 2П9М, pentode ГК-71.

For ultra-short waves: triodes ГУ-4, ГИ-3, ГУ-8, double beam tetrodes ГУ-29, ГУ-32, pentodes 2П29Л, 2П29П, 4П1Л, ГУ-15, ГУ-50, ГУ-13, ГУ-72, ГУ-80.

Any type of continuous wave transmitter puts out a steady flow of high-frequency energy. This high-frequency energy flow is called the *carrier wave*, or simply the *carrier*, because it carries a certain type of intelligence which a given radio transmitter must convey to radio receivers. Such intelligence may be represented by sound oscillations, these sound oscillations acting upon the carrier wave, by the process known as *modulation*. The most popular type of modulation

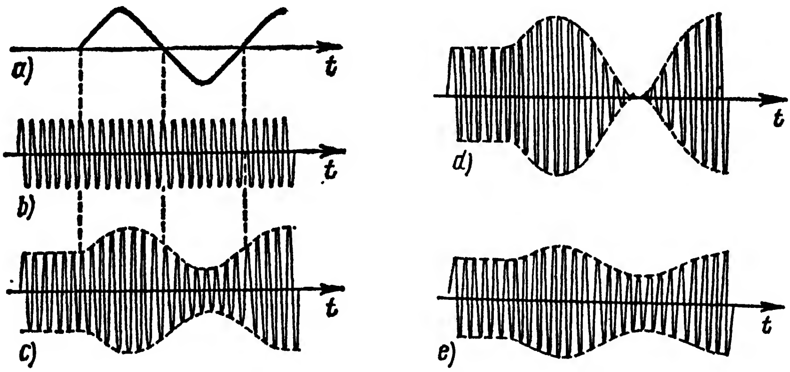


Fig. 188. Graphic representation of amplitude modulation: *a)* modulating low-frequency signal; *b)* high-frequency carrier; *c), d) and e)* modulated oscillations at respective modulation-factor values of 50 %, 100 % and 30 %

is called *amplitude modulation* in which the amplitude of the carrier is varied by sound oscillations. The variations of the carrier amplitude take place in exact conformity with the characteristics of the sound being transmitted.

Formulating the above-said, the following statement is in order: *modulation is the process of changing the amplitude of high-frequency oscillations in accordance with audio-frequency signals which are to be transmitted.* High-frequency oscillations whose amplitude varies in accordance with audio-frequency signals as a result of modulation are called *modulated oscillations. Modulators* is the name of electrical circuits performing the modulation process.

The simplest type of modulation is illustrated by connecting a carbon microphone into the aerial of a radio transmitter. When the microphone is spoken into, its resistance will vary in accordance with the vibration of its diaphragm. The variations of the microphone resistance will vary the amplitude of the high-frequency current flowing in the aerial, i.e., the microphone will directly modulate the carrier wave of the transmitter. This type of modulation, however, is no longer used by radio transmitters because the mod-

ulation is accompanied by a large loss of power in the microphone.

The modulation process can be shown graphically. Fig. 188*a* gives the curve of a sinusoidal audio-frequency oscillation to be transmitted by radio. Fig. 188*b* shows continuous unmodulated high-frequency oscillations — the carrier wave. And, finally, Fig. 188*c* shows modulated oscillations resulting from the application of the sinusoidal audio-frequency signal upon the high-frequency carrier. The amplitude changes of such modulated oscillation follow exactly all the characteristic features of the audio signal.

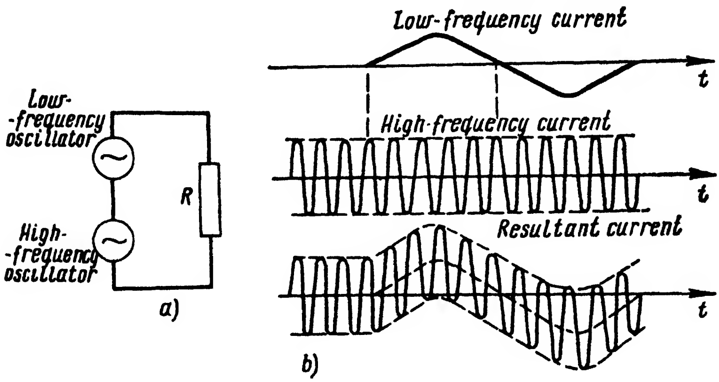


Fig. 189. The addition of high-frequency and low-frequency oscillations

Modulation should not be thought of as superimposition of low-frequency signals upon high-frequency oscillations. Likewise, it should not be regarded as the addition of low-frequency oscillations to high-frequency oscillations. And, again, it would be wrong to say that a modulated current is the sum of high- and low-frequency currents. Such addition of currents is not modulation and can be done by means of circuit shown in Fig. 189*a*. Here, a low-frequency oscillator is connected in series with a high-frequency generator and with certain load resistance *R*. Fig. 189*b* shows the addition of the currents of the two generators. The resultant current possesses a high-frequency component, whose amplitude remains constant. Evidently, this is not a case of modulation.

Depending upon the intensity of action of audio signals upon a carrier wave, the amplitude changes of the carrier will be greater or smaller. In this case, we speak of a greater or smaller *modulation factor* (sometimes called the *modulation percentage)*.

The modulation factor, denoted by the letter m, shows the maximum amplitude change of a modulated ossillation, when compared to the amplitude of the same oscillation before the modulating signal is applied (i.e., when compared to the amplitude of the carrier wave). The mod-

ulation factor is usually expressed in per cent; hence, the name — modulation percentage. Fig. 188 shows modulated oscillations with different values of modulation factor, the carrier being acted upon by a sinusoidal modulating signal. If during the modulation process the carrier amplitude changes to the extent of its full value, first decreasing to zero and then increasing to a doubled value, i.e., changing by 100% to both sides of its unmodulated condition, the modulation factor (percentage) is equal to 100%. This case is illustrated in Fig. 188*d*. If the carrier amplitude changes by 30% of its original value (Fig. 188*e*), the modulation factor is 30%, etc.

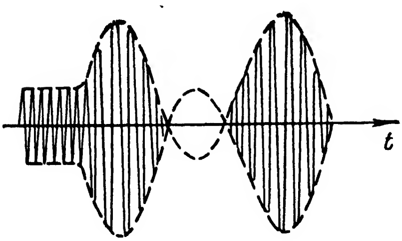


Fig. 190. A case of overmodulation

Cases are possible when the modulation factor exceeds 100% (Fig. 190). Such cases are known as *overmodulation*. Here, the carrier amplitude is increased by more than 100%. However, since it cannot decrease by more than 100%, a part of the oscillations will be cut off, which will result in distortion. Hence, the maximum value of modulation factor must not exceed 100%, if distortionless transmission is to be had. In actual transmission of speech or music by radio, the sounds being transmitted have different strength and, consequently, the modulation percentage does not remain constant (as it would in case of transmitting a steady sinusoidal sound). On strong modulating sounds the modulation percentage is increased, on weaker sounds — decreased. In order to avoid overmodulation, the transmitter must be so adjusted that m=100% only on the strongest sounds being transmitted, which means that the modulation percentage will be less than 100% when sounds of lower intensity modulate the carrier wave of the transmitter. As a rule, the carrier wave of an average radio station is modulated within the limits of 30-80%. However, since the communication radius depends upon the modulation percentage, some radio stations (particularly those transmitting speech only) resort to an increase of the modulation factor up to 85 and even 90%, allowing the distortion to take place on transmission of the strongest sounds. Such practice is, of course, to be avoided when the radio station transmits musical programmes, for which the value of m must be limited by 50-60%.

92. THE MAKE-UP OF MODULATED OSCILLATIONS

As shown in 1916 by the eminent Soviet scientist M.V. Shuleikin, modulated oscillations are a sum of several simple sinusoidal oscillations, differing in frequency from each other.

Oscillations, which are modulated by a sinusoidal sound, represent a sum of the following three continuous oscillations: the carrier oscillation which has a frequency equal to frequency f of the oscillation being modulated, and two other oscillations — the so-called lower and upper sidebands — whose frequencies are, respectively, lower and higher than

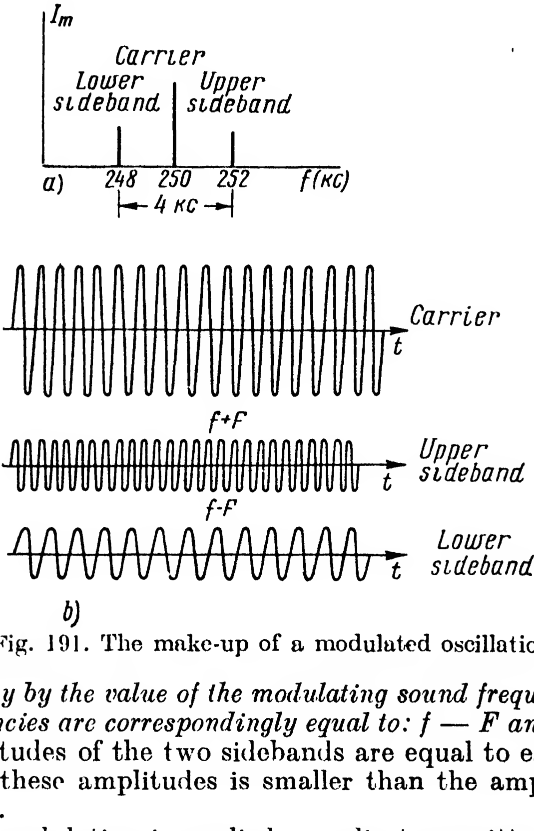


Fig. 191. The make-up of a modulated oscillation

the f frequency by the value of the modulating sound frequency F, i.e.— whose frequencies are correspondingly equal to: f — F and f + F.

The amplitudes of the two sidebands are equal to each other but each one of these amplitudes is smaller than the amplitude of the carrier wave.

When no modulation is applied, a radio transmitter radiates the carrier wave only. Sidebands are created as soon as modulation is applied. For instance, if the carrier frequency of a radio transmitter is given by $f=250$ kc and is modulated by a sound whose frequency (F) is equal to 2,000 cps (2 kc), then the transmitter radiates not only the 250-kc carrier but also the upper sideband, whose frequency is given by $f+F=252$ kc, and the lower sideband with a frequency of $f-F=248$ kc. Fig. 191*a* gives the spectrum of a modulated oscil-

lation, i.e., a distribution diagram of frequencies and amplitudes of components in an oscillation modulated by a simple sound. The components themselves are shown in Fig. 191*b*.

Let us examine a simple proof, which shows that sidebands exist in a modulated carrier. This proof will be useful to us in our later studies, when we come to the discussion of radio receivers. Let us try to add up all the three parts of modulated oscillations: the carrier oscillation and the two sidebands. Let us first add up the sidebands. When two continuous oscillations, whose frequencies are not equal, are added up, the so-called *beat frequencies*, or simply *beats*, are produced, such beats playing an important role in radio.

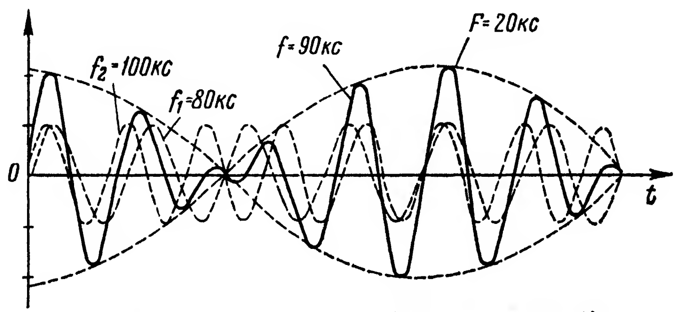


Fig. 192. The process of beat-frequency generation

The beats are represented by oscillations with periodically changing amplitude, resembling modulated oscillation to a certain extent. If the added-up oscillations have respective frequencies f_1 and f_2, the average frequency of the resultant oscillation is given by:

$$f = \frac{f_1 + f_2}{2}.$$

Here, the amplitude changes (pulsates) at a lower frequency, the latter referred to as the beat frequency F and equal to the difference of frequencies f_1 and f_2. Fig. 192 shows the addition of two oscillations with equal amplitudes but different frequencies: $f_1 = 80$ kc and $f_2 = 100$ kc (the ratio of frequencies is 4 to 5, while the ratio of periods is 5 to 4). As a result of such addition, a frequency oscillation is obtained, whose frequency is given by: $f = \frac{(80 + 100)}{2} = 90$ kc. The amplitude of this oscillation pulsates with a frequency $F = 100 - 80 = 20$ kc.

It should be noted that during the transition from one group of beats to another, an oscillation takes place, the period of such oscillation being twice as small as the period of the other oscillations. Thus, a complete oscillation takes place during one half-period and, therefore, the phase of the oscillation of each following group of beats is opposite in phase to the oscillation of a preceding group.

If now we add the beats to a carrier, whose frequency $f=90$ kc. then one group of beats will be added to the carrier while the other group of beats will be subtracted from it, inasmuch as the phases of beat groups will be in opposition and the phase of the carrier oscillations will be constant. As a result, a correct modulated oscillation will be obtained. This oscillation will have a high frequency $f=90$ kc and a modulation frequency $F=10$ kc (Fig. 193). But, then, this oscillation consists of the carrier oscillation and two sidebands with respective frequencies of 100 and 80 kc. Hence, we can consider, that analysis of modulated oscillation into its components is proven.

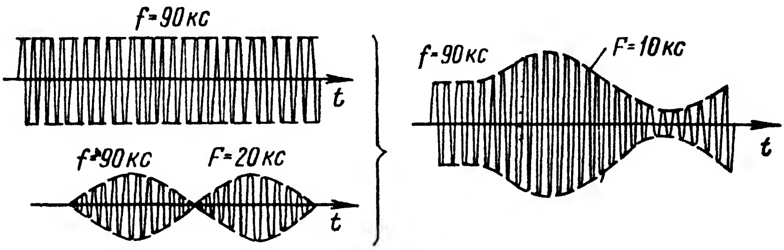


Fig. 193. Graphic representation of the make-up of modulated oscillations

Thus, a modulated oscillation can be regarded from two points of view: either as an oscillation of a definite frequency and varying amplitude, or else as a sum of the carrier oscillation and sideband oscillations with constant amplitudes but different frequencies Both these points of view are founded on the same physical phenomenon and do not contradict each other. They simply regard different sides of the same process. In our further discussion we shall, in each specific case, refer to that viewpoint which will allow us to get the simplest and clearest answer to any particular problem with which we might be dealing.

In a radio telephone transmission, the audio oscillations have a complex form. The frequency and amplitude of sounds are changing. Each complex sound is a sum of several oscillations which have different frequencies and amplitudes. It is these complex combinations of speech and music oscillations that perform the actual modulation; and, because of this, we have to deal not simply with two sidebands, but rather with two side frequency ranges. For instance, if a radio-telephone transmitter operates on a carrier frequency of 500 kc and the frequency range of the modulating sounds extends from 100 to 10,000 cps (0.1-10 kc), the upper sideband will contain frequencies from $500 + 0.1 = 500.1$ kc to $500+10=510$ kc, while the lower sideband will be taken up by

frequencies from $500-0.1=499.9$ kc to $500-10=490$ kc (Fig. 194). Hence, the actual band of frequencies radiated by such a transmitter will stretch from 490 to 510 kc.

The frequency bandwidth of oscillations radiated by a radio-telephone transmitter is equal to twice the value of the highest modulating frequency. In the example just given above the frequency bandwidth is equal to 20 kc (from 490 to 510 kc), because the highest audio frequency used to modulate the transmitter is 10 kc.

High-fidelity transmission of music and speech calls for the transmission of approximately such a band of frequencies. However, the great number of operating radio stations and the shortage of available waves in the broadcast band make it necessary to compress the frequency bandwidth transmitted by each individual radio station. At present, each broadcasting station is allowed to radiate a frequency bandwidth not in excess of 9 kc. This means, that a station is not permitted to transmit frequencies higher than 4,500 cps. Frequencies higher than 4,500 cps must be cut off at the station transmitter, so that the respective sidebands, radiated by the transmitter, would not interfere with the transmission of another station operating on an adjacent carrier frequency. If this rule is not observed, a radio receiver tuned to one of these stations will reproduce the interference in the form of squeals and whistles, which make the reception highly unpleasant.

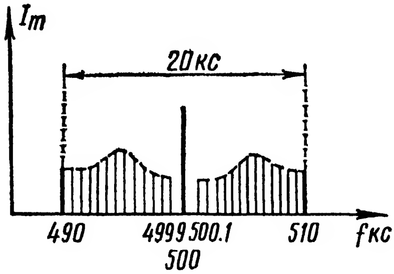


Fig. 194. Sideband frequencies on modulation

Under conditions of modulation, the amplitude of a carrier wave is steady, while the amplitude of sidebands varies, always remaining smaller than the carrier amplitude. The useful power in radio telephony is the power of the sidebands, and this power is smaller than the carrier power. Theory shows that even when $m=100\%$, the power included in both sidebands is equal only to one-third of the total radiated power, the other two-thirds of the power being included in the carrier. When $m=50\%$, the power of both sidebands is equal only to 0.1 of the total radiated power. Because of this, a radio-telephone station has a much smaller operating radius than a radio-telegraph station having similar power-output and identical conditions.

93. GRID MODULATION

There are several methods of amplitude modulation. In any one of these methods, the operating condition of a radio transmitter varies under the influence of audio frequency signals, as a result of which the amplitude of oscillations in the output circuit of the trans-

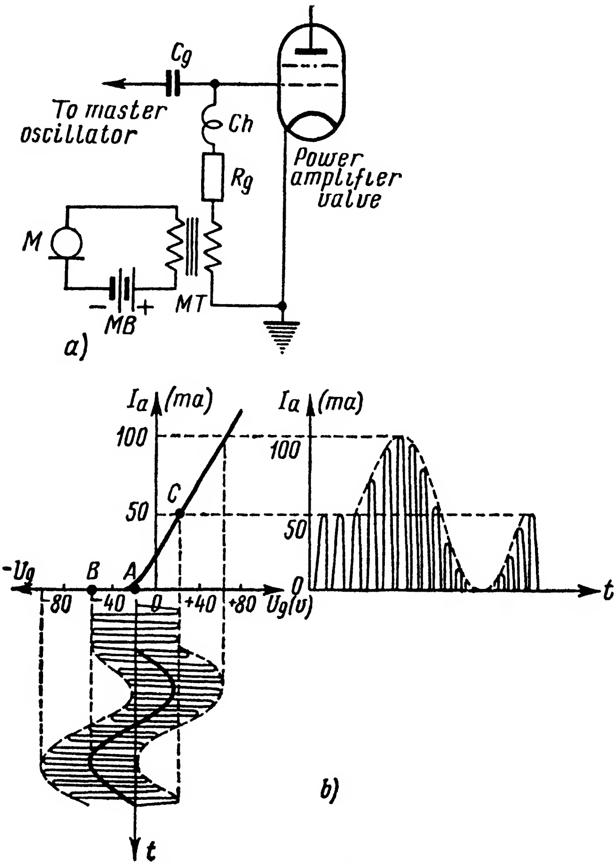


Fig. 195. A grid-modulated stage and graphic representation of grid-modulation process

mitter changes in accordance with the character of the sound under transmission. In order not to impair the frequency stability of the master oscillator, this stage is not modulated. The modulation is preferably applied to one of the power amplifier stages, usually to the final stage, but sometimes to an intermediate one.

The simplest practical method of modulation is *grid modulation.* In this method the grid bias voltage of the power amplifier valve is varied at audio frequency (Fig. 195*a*). This is obtained by connecting

the secondary winding of microphone transformer MT, in series with grid resistor R_g, into the circuit carrying the d.c. component of grid current. Thus, the control grid of the modulated power amplifier is fed not only with high-frequency excitation voltage from the master oscillator (or from a previous amplifier stage) and with d.c. bias voltage from the grid resistor, but is also fed with alternating audio-frequency voltage. The latter voltage changes the value of bias voltage and shifts the operating point on the characteristic of the modulated valve.

Fig. 195*b* gives a graphic representation of grid-modulated power amplifier valve operation, when the modulation factor is equal to 100%. In this case, the d.c. bias voltage is —20 v, the amplitude of the exciting high-frequency voltage is 40 v, and the amplitude of the modulating audio-frequency voltage is also 40 v. When no modulating signal is applied to the grid of the valve, the anode current pulse is equal to 50 ma. Modulation applied, the operating point shifts from position A to position C during each positive half-period of the modulating voltage, increasing the value of anode current pulses to 100 ma. During negative half-periods of the audio-frequency voltage, the value of bias voltage reaches —60 v, the operating point is shifted to position B, the anode current of the valve is cut off, and the value of anode current pulses becomes zero.

As a result of the described process, the high-frequency component of the anode current represents a modulated current, the modulation factor of which is 100%. This current feeds the tuned anode circuit of the amplifier, the oscillations in this circuit also assuming a modulated character. The modulation factor of the stage will be decreased with a decrease of the amplitude of the audio-frequency modulating voltage.

The initial operating condition of a modulated stage, before the modulation can be applied to it, must correspond to a class B or class C amplifier condition. Under this condition, the operating point of the valve characteristic will be located in position A. Modulation will not occur under class A condition, which can be easily proven by resorting to a graphic representation similar to that given in Fig. 195*b*. Under class A amplifier operating condition, the high-frequency component of the anode current will have a constant amplitude, i.e., will not be modulated. Besides this, when the stage is under its initial condition, the anode current pulses must have a value equal to one-half of the possible maximum value. This is necessary so that when $m=100\%$ the anode current pulses would reach double the initial value.

The operating condition with the maximum possible anode current pulses, giving the maximum power output, is called *the telegraph operating condition*. An increase of grid bias voltage halves the value of anode current pulses, thus halving the current in the tuned circuit

and in the aerial, and giving what is known as the *telephone operating condition*. Under this condition the power in the aerial is reduced four times, which considerably shortens the operating range of the radio transmitter.

A modulation characteristic is employed to determine the correct condition of modulation with the smallest distortion. Such a characteristic is plotted experimentally and represents the dependence of the high-frequency current in the aerial upon the bias voltage, while the excitation voltage is kept constant. An example of a modulation characteristic is given in Fig. 196, and the points of telegraph and telephone operating conditions are shown on the curve. The point of the telephone operating condition is located in the middle of the linear part of the characteristic. This point determines the value of correct bias voltage which must be applied to the grid of the modulated valve. The length of the linear portion of the curve determines the amplitude of the modulating voltage.

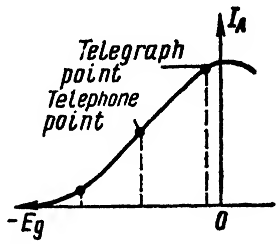


Fig. 196. The modulation characteristic for a case of grid modulation

The circuit given in Fig. 195 may be employed by low-power radio-telephone transmitters. The microphone transformer used in this circuit can be called a modulation transformer. The transformation ratio of this transformer may be from 1 : 10 to 1 : 100. If the heater of the valve in this circuit is fed with power from a d.c. supply, the same supply can be also used to feed the microphone circuit, and then the special microphone battery becomes unnecessary. If required, the supply voltage can be reduced by means of a dropping resistor connected in series with the microphone. However, modern radio-telephone transmitters seldom employ carbon microphones, preference generally being given to dynamic microphones which need no electric supply.

If the power of the grid-modulated stage is large, the modulator circuit must employ an additional audio amplifier. In this case, the microphone is connected through the microphone transformer to the input of such an amplifier, whose output circuit works into the primary winding of a special modulation transformer. The transformation ratio of this transformer is usually 1 : 1, the secondary winding of the transformer feeding the amplified modulating signal to the grid circuit of the valve being modulated.

A radio-telephone transmitter is checked for the presence of modulation by talking into the microphone, while watching the meters or other types of indicators. A hot-wire ammeter has a large thermal inertia and its readings will change but little on modulation. A better indicator is a small electric lamp connected into the aerial circuit. When the transmitter is modulated, the brightness of glow

given off by the lamp will vary in step with the modulating signal. A small neon lamp, because of its low inertia of ignition, can be used successfully instead of the incandescent lamp, if desired. The quality of modulation is usually checked (monitored) by listening in to the station on a crystal receiver.

94. ANODE MODULATION

When the power amplifier of a radio-telephone transmitter is modulated by varying the anode supply voltage at the frequency of the modulating audio voltage, the modulation is called anode modulation. As in the case of grid modulation, here, too, the modulated stage must be operated as a class B or class C amplifier. If the operating condition of the modulated power amplifier stage is correctly adjusted, the increase of the anode voltage on modulation will cause an increase of the amplitude of oscillations in the tuned circuit, while a decrease of the anode voltage will produce the opposite effect.

Theory shows that undistorted and high-percentage anode modulation is obtained only when the modulated stage operates under overexcited condition, i. e., when the anode current pulses are sharply non-sinusoidal and when spaces exist between them as a result of grid current increases (Fig. 171*c*).

In anode-modulated circuits, the modulating audio-frequency signal applied to the anode of the modulated stage must have a high power level. Because of this, a radio-telephone transmitter cannot be anode-modulated directly from a microphone, without any amplification — which was possible in the case of grid modulation. Even comparatively low-power anode-modulated transmitters require at least one stage of audio amplification, employing either a transformer-coupled or choke-coupled arrangement.

Speaking generally, a modulator, as a rule, is nothing but a low-frequency amplifier. The actual modulation process takes place not in the modulator but in the high-frequency amplifier stage which is being modulated. The purpose served by the modulator is that of amplifying the low-frequency signal, delivered by the microphone, until the signal becomes sufficiently strong to affect the operating condition of the modulated high-frequency stage and to secure a high modulation percentage.

Fig. 197*a* shows the circuit of anode modulation, where the modulator is transformer-coupled to the modulated stage and does not differ from a common transformer-coupled low-frequency amplifier stage. The secondary winding of modulation transformer T_2 is connected in series with the anode power supply of the modulated valve V_2. The alternating audio-frequency voltage developed in this winding is added to the d. c. voltage of the anode power

supply. This makes the anode voltage of valve V_2 pulsate at audio frequency, which produces the effect of modulation. In order to reduce the constant magnetisation of the modulation-transformer core, the following measures are resorted to: The currents drawn by the anode circuits of the modulator valve and of the modulated valve are made as equal as possible and the transformation ratio of

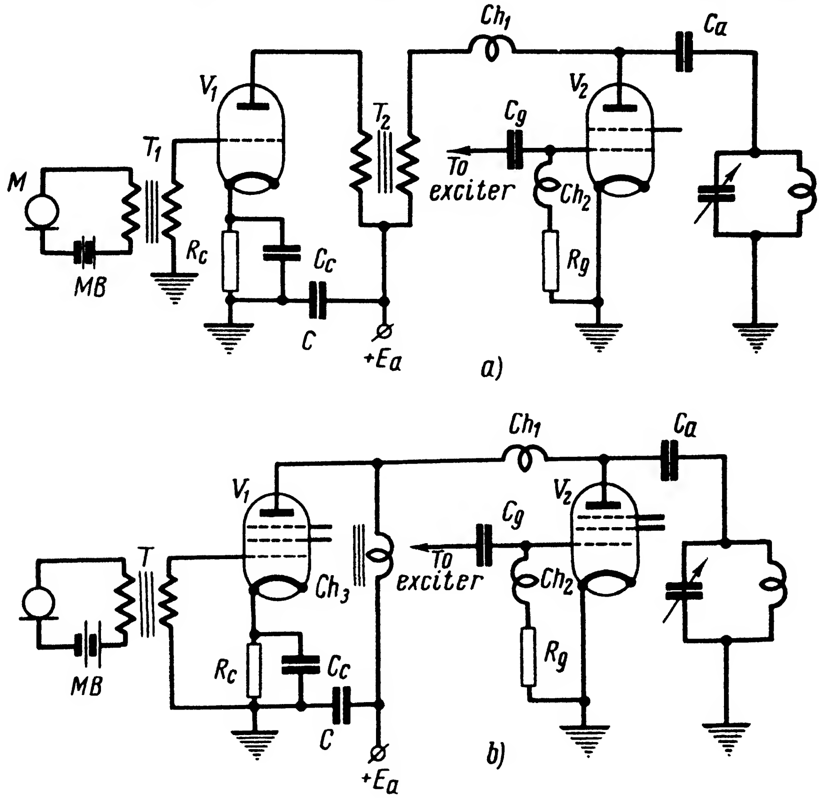


Fig. 197. Anode modulation circuits: *a)* a circuit with modulation transformer; *b)* a circuit with modulation choke

transformer T_2 is set to 1 : 1. This done, the ends of the two windings of T_2 are so connected that the currents flowing through these windings set up magnetic fluxes of opposite directions, the two fluxes, thus, mutually cancelling each other. In higher-power transmitters, an additional audio-frequency amplifier, sometimes referred to as *submodulator*, is provided between the microphone and the modulator stage. The modulator stage, in this case, employs a push-pull circuit in order to decrease non-linear distortion.

An alternative circuit of anode modulation is given in Fig. 197*b*. In this circuit, the modulator is represented by a choke-coupled

low-frequency amplifier. Modulation choke Ch_3 is connected in series with the common anode circuit of modulator valve V_1 and modulated valve V_2. A high audio-frequency voltage is built up across choke Ch_3 by the a. c. component of the anode current of modulator valve V_1. As a result, the anode voltage of modulated valve V_2 pulsates at audio frequency and modulation takes place.

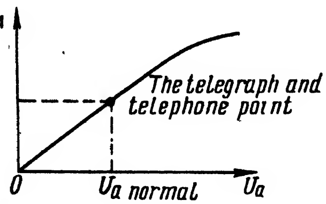


Fig. 198. The modulation characteristic for a case of anode modulation

If a high percentage of modulation is to be obtained, the power of the modulator valve must not be less than the power of the modulated valve. This is the disadvantage of anode modulation, as it is accompanied by high consumption of modulator anode and filament power. The advantage of anode modulation is seen in that the telephone-operating condition of an anode-modulated transmitter can be similar to its telegraph-operating condition. Hence, the output power of an anode-modulated transmitter is considerably greater than the output power of a grid-modulated transmitter using a similar type of modulated valve. When changing over from the telegraph to telephone operating condition, there is no need to reduce the output power of an anode-modulated transmitter—something which has to be done in all grid-modulated transmitters. Fig. 198 shows the modulation characteristic of an anode-modulated transmitter. On this characteristic, which represents the dependence of aerial current upon the anode voltage, may be seen the coincidence of the telephone-operating condition point with the point of the telegraph-operating condition.

Distortionless radio telephone transmission calls for high inductance value of modulation choke Ch_2. The core of this choke is provided with an air gap to prevent magnetisation of the core up to the point of saturation. Capacitor C, shunting the anode power supply, must have a large value of capacitance in order to pass the a. c. component of the audio frequency.

Anode modulation is successfully applied to high-frequency amplifier stages employing triodes, beam tetrodes and pentodes. It, however, cannot be applied to stages using ordinary tetrodes, because, at certain moments, the varying anode voltage will be lower than the screen-grid voltage and this would give rise to the dynatron effect, accompanied by strong distortion.

Grid modulation, or, alternatively, a *simultaneous anode and screen-grid modulation* may be applied to a tetrode valve. In the simultaneous anode and screen-grid modulation, shown in Fig. 199, the modulation choke (or transformer) is connected into the common

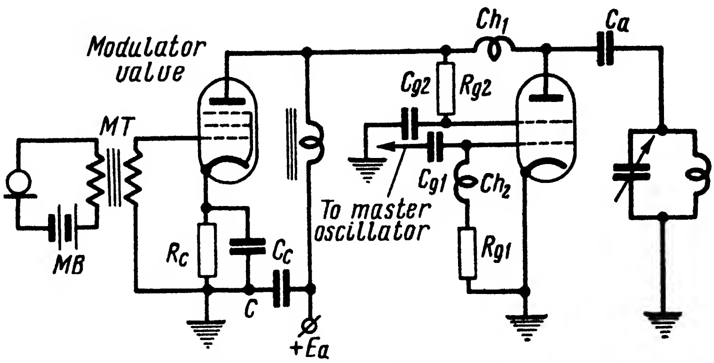


Fig. 199. A circuit of anode-screen grid modulation

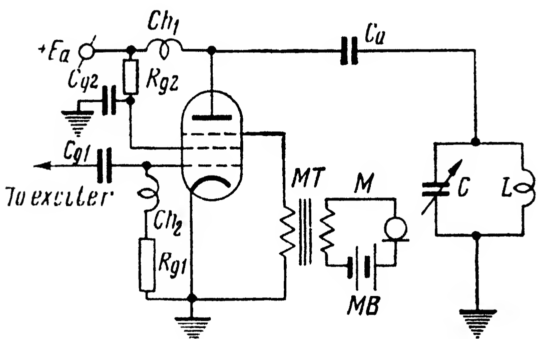


Fig. 200. Suppressor-grid modulation

anode and screen-grid circuit of the modulated high-frequency valve. In such a circuit, the anode and screen-grid voltages increase and decrease simultaneously, the anode voltage always remaining higher than the screen-grid voltage, thereby precluding the occurence of the dynatron effect. This type of simultaneous anode and screen-grid modulation is equally applicable to pentodes.

Suppressor-grid modulation can be also successfully applied to pentodes (Fig. 200). Such a circuit gives distortionless and high-

percentage modulation, and does this when the level of the modulating audio signal fed to the suppressor grid is comparatively low, even if the circuit employs a high-power pentode. In this circuit, a small negative bias voltage is usually applied to the suppressor grid.

96. FREQUENCY MODULATION

Frequency modulation, usually referred to simply as FM, is successfully employed by many modern radio communication and radio broadcasting systems.

This method of modulation is resorted to because of the following shortcomings of the usual amplitude modulation (AM).

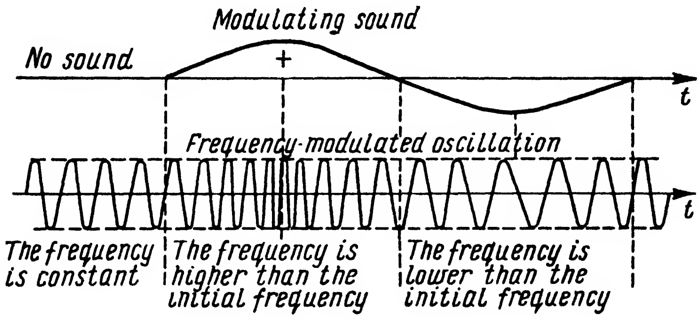


Fig. 201. Frequency modulation

AM transmitters are noted for low efficiency, as far as the utilisation of high-frequency power is concerned. As a result, the operating range of an AM radiotelephone transmitter is much shorter than the range of a radio-telegraph transmitter of similar power rating. Besides this, AM receivers reproduce various types of atmospheric and man-made electrical interference, making the reception of AM transmissions at times difficult and even impossible. Such interference, heard in radio receivers as rustling and crashing noises, is almost impossible to suppress. Frequency-modulation systems are free from these defects to a considerable degree, which accounts for the wide acceptance of FM.

In a frequency-modulated transmitter, the modulating audio-frequency signal varies the frequency and not the amplitude of the carrier wave. Fig. 201 gives a graphic representation of a modulating sinusoidal sound and shows how the sound varies the carrier frequency. As seen from the drawing, the carrier frequency gradually increases during one half-period of the audio oscillation and then returns to its initial value.

During the following half-period, the carrier frequency gradually decreases, reaches a certain minimum value, and then again returns to its initial value. The higher the amplitude of the modulating sound, the greater is the change of the carrier frequency. Radio broadcasting employs wideband frequency modulation, in which the frequency of the carrier deviates from its original value usually within $\pm$ 75-kc limits. Such wide frequency deviation is permissible only when the carrier itself has a sufficiently high frequency. Because of this, FM broadcasting has to employ ultra-short-wave ranges of frequencies higher than dozens of megacycles/sec.

The chief advantage of FM is seen in noise reduction, which gives a sharp improvement of radio reception. In comparison with an AM transmitter, a FM

transmitter makes a better use of high-frequency power, which increases the operating range of the FM radio station at a lower level of radiated power.

Narrow-band FM systems also exist. In such systems, the bandwidth of frequencies generated by the transmitter is about the same as the frequency bandwidth generated by an AM transmitter. Narrow-band FM systems do not, however, suppress interference as effectively as do the wideband systems.

In a FM transmitter, the audio-frequency signal must change, in one way or another, the frequency of the master oscillator. This can be done in various ways, one of which is shown in Fig. 202*a*. This is a very simple and well-operating circuit, in which frequency modulation is obtained by varying the so-called input dynamic capacitance of the modulator valve. The input capacitance C_i of this valve is connected, through capacitor C_1, in parallel with tuned circuit LC of the oscillator. The capacitance value of C_i depends upon grid bias voltage, which can be explained on the basis of the following simple considerations.

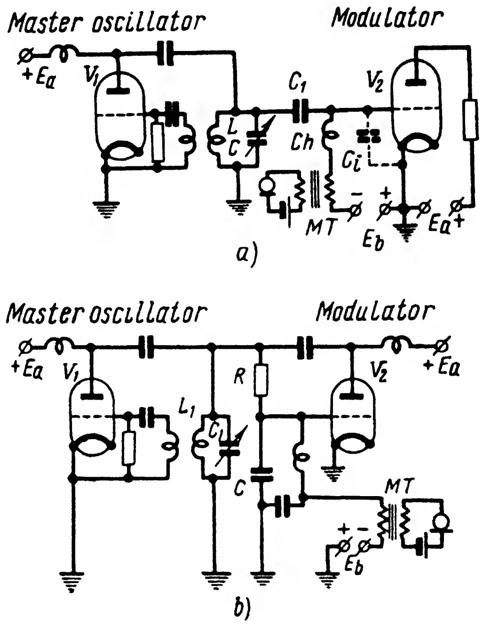


Fig. 202. Frequency-modulation circuits: *a)* a circuit functioning by virtue of modulator-valve input-capacitance change; *b)* a circuit employing a reactance valve

The master oscillator feeds high-frequency voltage to the grid of the modulator valve. The amplitude of this voltage is constant and its frequency changes only insignificantly during the process of modulation. For instance, in modern FM radio broadcasting transmitters, operating on metric waves, i.e., on frequencies of several dozens of megacycles, the modulation changes the frequency by only ± 75 kc. Hence, the frequency of grid voltage may be considered constant. Then, the input capacitance of the valve will be proportional to the value of capacitive grid current appearing because of the presence of this input capacitance.

If the input capacitance were equal to zero, there would not be any capacitive grid current. The larger the input capacitance, the greater will be the current flowing through it. This current comprises two currents. One of these currents flows through the grid-cathode capacitance of the valve, while the other current passes through the anode-grid capacitance. The value of the latter current is proportional to the alternating voltage between the anode and grid, this voltage making the current pass through the given capacitance. If the grid bias voltage is increased, the operating point will be moved along the valve characteristic into the region where the slope is less. This will reduce the amplification of the stage. The a.c. voltage between the anode and grid will also be reduced, which, in its turn, will decrease the capacitive current flowing through capacitance C_{ag}. But this will bring about a reduction of the total capacitive current in the grid circuit, which is equivalent to a reduction of the input capacitance. Apparently, the process will be reversed if the grid bias voltage is reduced, and the input capacitance will be increased.

Microphone transformer *MT* is connected into the grid circuit of valve V_2. When audio-frequency voltage is fed through this transformer, the bias of valve

V_2 will vary. Hence, the input capacitance of the modulator valve will change and, consequently, the frequency of the master oscillator will also vary.

Another popular circuit of a FM transmitter employs the so-called reactance valve. This circuit is shown in Fig. 202*b*. Here, the internal resistance of modulator valve V_2, i.e., the impedance between the anode and cathode, has an inductive character and its value depends upon the value of grid bias voltage. This internal resistance is equivalent to a certain inductive reactance and is in parallel with the tuned circuit of the master oscillator. In other words, it is a part of this tuned circuit and partly determines the frequency of the given circuit. If an audio-frequency voltage is applied to the grid of the modulator valve, the reactive internal resistance of the valve will vary, which will cause a variation of the master oscillator frequency.

The reactive character of the internal resistance of the modulator valve is explained as follows. High-frequency alternating voltage from tuned circuit C_1L_1 of the oscillator is fed to the anode of the modulator valve and, simultaneously, to its grid through a divider comprised of resistor R and capacitor C. The resistance value of resistor R is made considerably larger than the capacitive reactance of capacitor C. Because of this, the phase of the current flowing through the RC circuit practically coincides with the phase of voltage supplied from the oscillator, i.e., coincides with the phase of the a.c. voltage at the anode of the modulator valve. The voltage across capacitor C and, hence, the voltage at the grid of the modulator valve, will lag behind the current by 90°. Under the influence of this voltage, an alternating anode current will be set up in the valve. The phase of this current may be considered coincident with the phase of the voltage at the grid. But this means that the current will lag in phase behind the anode voltage by 90°. It, then, follows that the internal resistance of the valve really has an inductive character, i.e., the space between the anode and the cathode of the valve behaves as an inductance. The value of this inductance depends upon the mutual conductance of the valve. If, for instance, the mutual conductance is reduced, then the a.c. anode current value will be lowered at the initial voltage values. This is equivalent to an increase of the internal resistance. Or, in other words, this is the same as an increase of the inductance equivalent to the valve. When the grid of the modulator valve is fed with an audio-frequency voltage, the grid bias and the mutual conductance of the valve will be changing with the frequency of this voltage. As a result, the equivalent inductance of the valve will be changing, which will cause a change of the oscillator frequency.

If R and C of the divider feeding the modulator valve grid with voltage are made to change places, and if the value of R is made much smaller than the capacitive reactance of C, then, after repeating all the above considerations, it is easy to come to the conclusion that the valve is equivalent to a certain capacitance, whose value depends upon the grid bias voltage. Thus, in this case the principle of modulation remains unchanged. It is also possible to replace capacitor C by an inductance coil in the modulator circuit employing the reactance valve.

Another advantage of frequency modulation over amplitude modulation is seen in the following. Since the frequency modulation takes place in a low-power master oscillator, a FM transmitter needs no high-power modulator. The transmitter, in this case, employs the modulator only for the purpose of affecting the parameters of the tuned circuit of the master oscillator, thus varying the oscillator frequency.

Additional information on FM will be given under Chapter IX in the discussion on the reception of frequency-modulated signals.

97. QUESTIONS AND PROBLEMS

1. What is the difference between self-excited and separately-excited oscillators?
2. What is a feedback?
3. What are the conditions of self-excitation in a valve oscillator?
4. $L = 120\ \mu h$ and $C = 200$ pf in the tuned circuit of a valve oscillator. Determine the oscillation frequency of the given oscillator.
5. Why is it that too strong a feedback is not advantageous in a valve oscillator?
6. Saturation current I_{sat} of a valve is equal to 200 ma. Anode voltage $U_a = 450$ v. What will be the useful energy output of this valve when used as an oscillator?
7. Anode voltage U_a of a valve oscillator is equal to 750 v, and the d.c. component I_a of the anode current is 40 ma. The useful power output $P_K = 18$ w. Find the value of anode dissipation P_a and the efficiency of the oscillator.
8. What is the purpose served by grid bias voltage in a valve oscillator?
9. Is it always possible to obtain the following value of useful power from a valve oscillator: $P_K = 0.2 I_{sat} U_a$?
10. The resistance (impedance) of the tuned circuit of a valve oscillator is given as $R_e = 10{,}000$ ohms at resonance. The a.c. voltage built across the tuned circuit has an amplitude given as $U_{ma} = 400$ volts. Find power P_K in the given tuned circuit.
11. What will happen in a valve oscillator if its grid capacitor is punctured during the operation of the stage?
12. Alternating voltage applied to the grid of a valve has the following values of amplitude and frequency: $U_{mg} = 100$ v, $f = 3{,}000$ kc. Grid resistor R_g, connected between the grid and cathode, has a value of 10,000 ohms. What is the loss of high-frequency power in this resistor?

By how many times will this loss be reduced if choke L, having a value of 2 mh, is connected in series with resistor R_g?

13. Why does the bias voltage, developed in a grid resistor, drop to zero if a valve oscillator, employing such resistor, stops functioning?
14. Draw the circuit diagram of a valve oscillator employing an inductive feedback, parallel anode feed, a tuned grid circuit, and an ohmic resistor in place of anode choke. Why is it that such a circuit is applicable only to low-power valve oscillators?
15. Draw the circuit of a tuned-anode tuned-grid valve oscillator with parallel anode feed.
16. Draw the circuit of a push-pull valve oscillator with capacitive feedback and parallel anode feed.
17. What will happen in the circuit of a valve oscillator with parallel anode feed, if the blocking capacitor is punctured?
18. Why is it that an oscillator with parallel anode feed cannot function without a choke in the anode circuit?
19. Why is it convenient to couple a radio transmitter to its associated aerial by a variable inductance?
20. Is it true that the main function of a power amplifier consists of amplification of oscillations generated by the master oscillator?
21. Why cannot the aerial tuning capacitor of a radio transmitter be ganged on the same shaft with the capacitors used to tune the closed oscillatory circuits of the transmitter?
22. The tuned circuit of a power amplifier has the following parameters: $L = 80\ \mu h$, $C = 150$ pf. Determine the oscillation frequency of the radio transmitter employing such amplifier.
23. What is the purpose served by an aerial indicator?
24. Does the frequency of a radio transmitter depend upon the parameters of its aerial?

25. Why is it necessary to tune the aerial of a radio transmitter to the frequency of the master oscillator?

26. What is the undesirable effect produced by anode-grid capacitance in a power amplifier?

27. Draw the circuit diagram of an electron-coupled ratio transmitter, in which the master oscillator employs additional inductive feedback.

28. Why is frequency multiplication employed in valve oscillators?

29. How is it possible to convert an ordinary power amplifier into a frequency doubler?

30. What are the main reasons of frequency instability in a radio transmitter?

31. Why will the frequency of a master oscillator stage change, if the valve employed by such stage is replaced by another valve?

32. What is the thickness of a quartz plate designed to oscillate with wavelength $\lambda = 40$ m?

33. What is the peculiarity of zero-angle cut crystals?

34. Draw circuit diagrams of crystal-controlled oscillators employing auxiliary autotransformer and capacitive types of feedback.

35. Why is it not advisable to connect a telegraph key into one of the circuits of the master oscillator?

36. Is it a good practice to connect a telegraph key into the filament circuit of a power amplifier?

37. What are the negative signals?

38. Why is it not advisable to use triodes in power amplifiers?

39. Draw the circuit diagram of a separately-excited radio transmitter, in which the master oscillator employs an electron-coupled pentode, and a pentode valve is also used in the power amplifier stage. Use your own judgement when selecting all the remaining specifications of the transmitter.

40. Where is the question of stage efficiency of greater importance: in a 5-watt master oscillator or in a 10-kilowatt final amplifier stage of a radio transmitter?

41. Can a frequency doubler function under the conditions of a class A amplifier?

42. Are there any self-excited oscillators which use no feedback?

43. What are the advantages of push-pull valve oscillators?

44. Why is it not advisable, in the case of a self-excited valve oscillator, to use bias voltage developed by a separate supply source or to employ automatic biasing circuit, in which the bias voltage is provided by the anode current drop?

45. The amplitude of high-frequency current in the aerial of a radio transmitter is equal to 20 amperes when the carrier is not modulated. When modulation is applied, the amplitude at certain moments increases to 32 amperes, dropping to 8 amperes at other moments. What is the modulation factor?

46. Would it be correct to say that modulation takes place in the modulator of a radio telephone transmitter?

47. Is the modulation factor a constant value in a radio transmission?

48. The frequency of oscillations generated by a radio transmitter is 600 kc. These oscillations are modulated by a 200-cps sinusoidal sound. What is the frequency make-up of the modulated oscillations in the given case?

49. What are the beat frequencies?

50. What is the frequency bandwidth of a radio telephone transmitter, if the highest frequency of the modulating sounds is 2,500 cps?

51. What is the difference between frequency-modulated and amplitude-modulated oscillations?

52. The amplitude of aerial current increases to its double value and decreases to zero when the modulation factor is equal to 100%. How does the aerial power change under such conditions?

53. Draw the circuit diagram of a radio transmitter employing control grid modulation and possessing the following features: a) the master oscillator employs an autotransformer type of feedback, a pentode and parallel anode feed; b) the power amplifier also uses a pentode and parallel anode feed; c) the modu-

lator consists of a single choke-coupled triode stage, i.e., operates without a modulation transformer.

54. Anode modulation can be effected by a circuit in which the modulation choke is replaced by a common ohmic resistor. However, such a circuit will have a serious shortcoming. What is this shortcoming?

55. Inductance L of a modulation choke is equal to 40 henries. Find the inductive reactance of the choke for frequencies of 50 and 3,200 cps.

56. Capacitor shunting the anode battery in a radio transmitter with anode modulation has a capacitance value C of 5,000 pf. Is this capacitance sufficiently high to allow the capacitor to pass the low-frequency component of the anode current?

57. Draw the circuit diagram of a radio transmitter employing electron coupling and suppressor-grid modulation.

58. What connections must be switched in the circuit given in Fig. 199 in order to obtain the true anode modulation in place of the anode-screen grid modulation?

59. In what stage of a radio transmitter does frequency modulation take place?

60. Draw the circuit diagram of a radio transmitter employing anode modulation, a modulation transformer and a push-pull modulator stage using triodes.

61. In what parts of a radio-telephone transmitter do the frequency distortion and non-linear distortion take place? Explain in detail.

CHAPTER IX

RADIO RECEIVERS

98. GENERAL DEFINITIONS

The general principle of operation of a radio receiver may be explained as follows. When a radio wave reaches the aerial wire of a radio receiver, it sets up an alternating e.m.f. in such a wire. This e.m.f. generates an alternating current in the aerial, the frequency of this current being equal to the frequency of the received wave, i. e., equal to the frequency of the current flowing in the aerial of the transmitter. It should be understood, however, that the oscillatory power in the aerial is usually negligibly small and must be amplified before it becomes of any practical use. Such amplification is performed by adjusting the tuned circuits of a radio receiver to resonance and also by resorting to the amplification properties of electron valves. Generally speaking, there are two kinds of radio receivers, as defined below — crystal receivers and valve receivers.

The *crystal receiver* is the simplest type of radio receiver, employing no electron valves and no electric power supplies, but offering only weak earphone reception of powerful radio stations, situated a short distance away from the reception point.

The *valve receiver* is a radio receiver employing electron valves and associated power supplies and, thereby, possessing high amplification. Such receivers are capable of amplifying even the weak signals of low-power and remote radio stations, received by the aerial, to a sufficient level for reproduction of such signals by earphones and even loudspeakers. Semi-conductor receivers, in which transistors take the place of valves, have the same properties as valve receivers and are discussed in detail under Chapter XI.

The amplification of signals picked up by the aerial is the first task of a radio receiver. It often happens that the gain provided by a single stage of amplification is insufficient. Because of this, oscillations amplified by the first stage of a radio receiver are fed to the second stage for the purpose of further amplification, then to the third stage, etc., until they are amplified to a required level. A radio receiver can employ 4, 5, 6 and even more amplification

stages. The total gain of the whole receiver is equal to the product of individual amplification factors of the stages. This gain can come up to several millions.

The second task of a radio receiver is the selection of a required radio station out of the numerous carriers reaching the aerial. It should be noted that a great number of carrier waves radiated by various radio stations reach every receiving aerial. Tuned circuits, used by every radio receiver and adjusted to resonance with the frequency of the desired radio station, amplify only the signals radiated by such a station. This important property of a radio receiver is called *selectivity*. If a radio receiver had no selectivity, all the signals received from various radio stations would be mixed in it and the reception of any particular radio station would become impossible.

The third problem solved by a radio receiver is that of detection, i. e., of conversion of high-frequency oscillations into low-frequency oscillations, the latter being a replica of the signal modulating the carrier wave of the station to which the receiver is tuned. The low-frequency signals (speech, music, etc.), thus reproduced by a radio receiver as a result of detection, are usually again subjected to amplification and are fed to earphones or a loudspeaker, where they are converted into sound.

There are two types of valve radio receivers. If frequency conversion takes place but once in a radio receiver, i. e., if high-frequency oscillations are directly converted into audio-frequency oscillations, the receiver is called a *straight-amplification receiver*. Radio receivers of this type are simple in design and were widely used in the past. But, at present, such receivers have been almost completely supplanted by another type of receivers known as *superheterodyne receivers*. A double frequency conversion takes place in a superheterodyne receiver. The design of such receivers is more complex than that of straight amplification receivers, but a superheterodyne receiver gives a higher amplification and a higher selectivity than a receiver of the other type (see Sec. 106).

In superheterodyne radio receivers, high-frequency oscillations, picked up by the receiving aerial, are converted in a special stage into the oscillations of another frequency, much higher than the audio frequency. These oscillations, known as intermediate-frequency oscillations and having a constant frequency for the given receiver, are then subjected to the basic amplification, during which the required selectivity is obtained. Following this, the amplified signals are subjected to detection, i. e., the intermediate-frequency signals are converted into audio-frequency signals.

This is the way in which the *double conversion*, mentioned above, is performed in a superheterodyne radio receiver. As already stated, it is because of the conversion of high frequency into intermediate frequency that high amplification and good selectivity are realised in a superheterodyne receiver, which is an invaluable feature in the reception of weak signals.

99. THE BASIC PARAMETERS OF RADIO RECEIVERS

A modern radio receiver must be capable of giving a good audibility on the reception of weak signals in a required frequency range. The receiver must also have a good selectivity and should not introduce noticeable distortion. The stability of the receiver is of great importance, too. The tuning of the receiver to a required wave should not change automatically, otherwise the signal will cease to be received. And even better, — in many cases, the receiver tuned to a definite wave must secure the stability of communication without any additional trimming of its tuned circuits; that is to say, the tuning should automatically follow slight drifts of the signal. Each receiver is characterised by the basic values given below.

Output voltage and output power. The final stage of a radio receiver feeds a certain level of low-frequency power to a pair of earphones or to a loudspeaker. This power is known as the *output power.* The voltage developed across the earphones and the loudspeaker is called the *output voltage.* A good earphone reception is secured when the output power level ranges within 10-20 milliwatts. Under such condition, a voltage of 15-20 v must be built up across high-impedance earphones. In case of low-impedance earphones, this voltage must be equal to 2-3 v. A much higher power-output level must be developed at the output of a radio receiver feeding a loudspeaker, this level reaching several watts in some receivers.

Sensitivity. The ability of a radio receiver to pick up and reproduce weak radio signals is called *sensitivity.* The sensitivity of a radio receiver is determined by the value of high-frequency voltage which must be fed to its input circuit (between the aerial-earth terminals) in order to secure a normal output power, i. e., to secure a normal reception. *The lower* is such input voltage necessary for the normal reception, *the higher* is the receiver sensitivity. The sensitivity of modern radio receivers ranges from several microvolts to several millivolts and depends upon the number of amplification stages and upon their quality. It is impracticable, however, to employ too great a number of stages, as this would inevitably increase the distortion, and valve noise, thus making a stable operation of the radio receiver difficult to attain.

Selectivity. The ability of a radio receiver to separate the signal of a required radio station from the signals of unwanted stations, operating on other frequencies, is called *selectivity.* In other words, *the selectivity of a radio receiver is its ability of receiving radio signals within a comparatively narrow frequency band.*

The selectivity of radio receivers is of paramount importance in our days, when a great number of radio stations, in many cases operating on nearly equal frequencies, are on the air. The higher the selectivity of a radio receiver, i.e., the narrower the frequency

band the receiver passes, the lower will be the interference set up in the given receiver by the signals of other radio stations.

The selectivity of radio receivers is usually shown by their resonance curves, known as the *selectivity curves*. When a selectivity curve is plotted, the selective properties of the whole receiver must be taken into consideration. Because of this, such a curve usually depicts a graphic relationship of output voltage U_0 and frequency f at the input of the receiver. The selectivity curve of a good radio receiver is shown in Fig. 203*a*. As follows from the study of this curve, the receiver is capable of receiving signals in a narrow frequen-

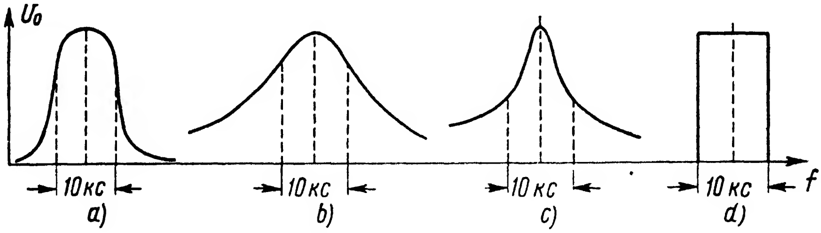


Fig. 203. Various shapes of receiver selectivity curves

cy band, which, as an example, is taken equal to 10 kc. The signals of other radio stations, possessing frequencies which do not fall within this band, are attenuated by many times. Some examples of poor-selectivity curves are shown in Fig. 203 *b* and *c*. Receivers possessing such selectivity curves will strongly reproduce the interference created by the signals of other radio stations operating on adjacent frequency channels. If the selectivity of a radio receiver is excessively high, the frequency band passed by the receiver becomes very narrow. This is advantageous on the reception of radio telegraph signals but creates considerable frequency distortion on the reception of radio broadcast programmes. Thus, each type of radio reception (radio telegraph, commercial radio telephone, radio broadcast) requires a different bandwidth of frequencies being passed.

Selectivity is frequently expressed by the attenuation of a received signal when the radio receiver is detuned from the given signal by a definite number of kilocycles. A selectivity curve helps to determine by how much the output voltage decreases at a certain value of detuning. When the selectivity of a radio receiver is good, a 5-kc detuning must attenuate the signal by at least 100 times.

The selectivity of a radio receiver depends upon the number and the quality of tuned circuits employed by the receiver. The greater the number of tuned circuits adjusted to resonance in a radio receiver, and the higher the quality of such tuned circuits (i.e., the lower is their damping), the higher is the selectivity of such a receiver.

However, only superheterodyne receivers are capable of making practical use of a large number of tuned circuits.

The quality of reproduction. The lower the distortion introduced by a radio receiver, the higher is the quality of reproduction of such a receiver. Frequency distortion and non-linear distortion, arising in low-frequency amplifiers, were discussed in Chapter VII. In a radio receiver, the distortion arises not only due to the distortion in the low-frequency amplifier, but also due to the resonant properties of the tuned circuits.

The broader the frequency band passed by a radio receiver, the more natural will sound the music and speech reproduced by such a receiver. Various types of intelligence transmitted by radio require the following bandwidths: speech — from 200 to 2,000 cps; music — from 100 to 5,000 cps; high-fidelity musical programmes — from 50 to 10,000 cps. It should be noted, however, that the requirement of broad-band transmission conflicts with the requirement of high selectivity. Higher selectivity leads to narrowing down the frequency bandwidth passed by a radio receiver and impairs the naturalness of sounds being reproduced. Such selectivity results in a particularly noticeable attenuation of the higher-frequency sounds, because, as a result of the sharp shape of the resonance curve, the sideband frequencies, which are cut off, are the ones which are most remote from the carrier frequency, and which represent high-frequency sounds.

An ideal resonance curve must have a rectangular shape, shown in Fig. 203*d*. A radio receiver possessing such a curve would pass equally well all the frequencies radiated by the radio station to which the receiver is tuned. Such an ideal resonance curve is, however, but an assumption. Practically obtainable curves, given in Fig. 203 *a*, *b* and *c*, produce a certain attenuation of the sidebands. The quality of sounds reproduced by a radio receiver can be improved by slight detuning of the tuned circuits of the receiver. When such detuning is present, the maximum point of the resonance curve will no longer correspond to the carrier frequency but rather to a certain sideband frequency. If the receiver is tuned exactly to the carrier frequency of a radio station, it will sound muted and there will be a deficiency of high-frequency tones in its output. The above-mentioned detuning from the carrier will restore the missing high-frequency tones and the reproduction of the programme will sound natural. When band filters, described in Sec. 108 are included in a radio receiver, the set will have a good selectivity and will pass a sufficiently broad band of frequencies.

Wavelength range. A radio receiver must be capable of being tuned to any wave within the wavelength range for which it is designed; and it is desirable that the sensitivity and selectivity of the receiver remain constant at all points of such a range. Radio broadcast receivers are usually designed to cover the medium-wave range

(200-600 m; 1,500-500 kc), the long-wave range (750-2,000 m; 400-150 kc), and sometimes that part of the short-wave range which is allocated to broadcasting (15-50 m; 20-6 mc). Radio receivers intended for special services are designed to operate on other wavelength ranges.

Apart from the wavelength coverage requirement, radio receivers also have to meet the requirements of stability and dependability of operation, economy of power consumption, convenience and simplicity of control, durability, easy access to the chassis wiring, etc.

100. DETECTION

A radio-telephone transmitter radiates modulated waves which, upon reaching the aerials of radio receivers, set up modulated high-frequency oscillations in such aerials. The *conversion of modulated high-frequency signals into audio-frequency signals is known as detection.*

If a modulated high-frequency current is passed through an earphone, no sound of any kind will be reproduced by the earphone because its diaphragm has a mechanical inertia and will remain stationary, not being able to follow such rapid oscillations. In other words, the modulated high-frequency current does not contain any audio frequencies capable of vibrating the diaphragm.*

The process of detection is performed with the help of special devices, called detectors, and possessing asymmetrical conductivity. A detector offers but a low resistance to currents flowing in one direction, and a much higher resistance to currents flowing in the opposite direction. Some types of detectors do not pass any currents at all in the opposite direction.

Modulated oscillations are no longer of symmetrical shape, once they have been passed by a detector; i.e., the adjacent positive and negative current half-waves will no longer have equal amplitudes after the detection; some half-waves, for instance the negative ones, become of a much smaller amplitude than the other half-waves (the positive ones, in the given case). In some cases, the detector will completely cut off one-half of the wave. As a result of this, an asymmetrical alternating modulated current (i.e., a pulsating current) is obtained at the output of the detector.

Fig. 204*a* gives a graphic representation of the modulated voltage applied to a detector, while Fig. 204*b* gives a similar representation of the pulsating current at the output of the detector. The current curve, shown in Fig. 204*b*, pertains to the case when the detector

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* Even if the diaphragm could follow high-frequency oscillations, it still would not reproduce any sounds, because high-frequency oscillations are very much beyond the audible frequency range.

completely cuts off the negative half-waves, not allowing them to pass through it at all.

The current flowing through the detector is a sum of the following three currents: the modulated high-frequency current, direct current, and audio-frequency alternating current. The sum of the direct current and of the audio-frequency current constitutes the current which pulsates at audio frequency. This current is shown by the broken line in Fig. 204*b*.

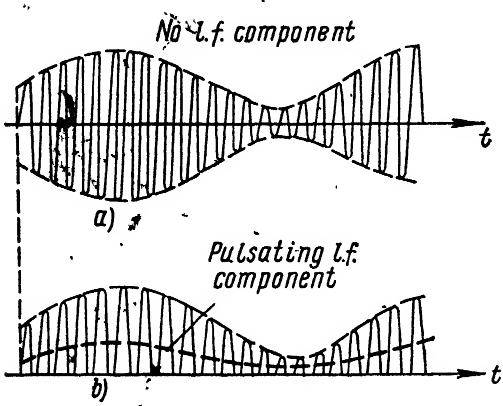

Fig. 204. Curves representing the process of detection

Thus, the detection process has given birth to a d.c. component and a low-frequency (audio-frequency) component, which were not present in the original modulated oscillation. It is this latter component — the audio-frequency component— that actuates the diaphragm of an earphone, the latter producing the original modulating sound. Neither the d.c. component nor the high-frequency component play any direct role in the reproduction of this sound.

To summarise the above-said, the following statement is in order: *during the process of detection, modulated oscillations act upon a detector noted for its asymmetrical conductivity, as a result of which there appears an audio-frequency current at the output of the detector, this current actuating earphones or a loudspeaker.*

101. CRYSTAL RECEIVERS

The crystal receiver is the simplest radio receiver. The detector used in such a receiver is a crystal or semi-conductor diode, in which a conducting crystal and metal are brought into contact. A detailed study of crystal detectors and of the processes taking place in them is given under Chapter XI.

In the case of a crystal receiver, a good reception is obtained only when the crystal-metal contact possesses a sufficient "sensitivity", i.e., a good detection effect. In old-type crystal receivers, the crystal was inserted into a special metal cap and was secured in it by means of a threaded lid or by a fusible metal. The most popular type of crystal was the *galena crystal* (galena is the name of the artificial sulphurous lead, which is a compound of lead and sulphur).

A 0.15-0.25 mm steel or copper wire was brought into contact with the crystal, the contacting end of the wire sharpened, for instance, by a diagonal cut. The wire was attached to a little lever and moved over the crystal surface until the most sensitive detection point was located, in which the wire (commonly known as a "catwhisker") was left. The "catwhisker" type of receiver was very unstable in operation, because the slightest push could displace the wire tip from the sensitive point on the crystal. This weakened the reception and made it necessary to move the "catwhisker" over the crystal again, looking for a new sensitive point. As a result, the crystal surface soon became scratched up and eventually damaged by the "catwhisker".

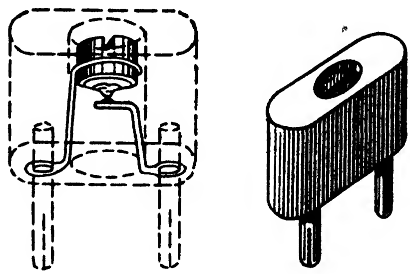

Fig. 205. A fixed-point silicon detector

Germanium and silicon crystals with a fixed sensitive point were developed ~~later~~ and are manufactured up to this date. Fig. 205 gives the construction of a silicon detector. The detector is placed in an ordinary electric lamp plug, in which a silicon crystal is brought into permanent contact with a small spring made of sheet-bronze or brass. If, as the time passes on, the sensitivity of such a detector becomes lowered, a new point of good contact may be located by turning the crystal-holding cap with a screw driver.

A complete circuit of the simplest crystal detector is given in Fig. 206*a*. Here, the tuned aerial circuit is comprised of coil *L* and capacitor *C*, the latter made variable and used to tune the aerial circuit to resonance with the frequency of a radio station being received. The modulated high-frequency voltage, built up across coil *L*, acts upon the detector circuit consisting of detector *D* and a pair of earphones *T*. The detector is used to rectify the high-frequency current, the resulting audio-frequency component actuating the diaphragm of the earphones.

The receiver shown in Fig. 206*a* employs a circuit known as a ***simple circuit***, because in this case the aerial constitutes a part of the tuned circuit of the radio receiver. The aerial itself possesses a certain natural wavelength, which is determined by the length of the aerial wire. Coil *L*, connected in the aerial, lengthens this natural wavelength. Conversely, the series-connected capacitor *C* shortens the wavelength, because the total capacitance is decreased with such a connection. On the other hand, if capacitor *C* were connected in parallel with the capacitance of the aerial, an increase of the total capacitance would have taken place and the wavelength would have been lengthened. In order to secure the coverage of a wide frequency range, some detector receivers resort to a switching arrange-

ment which permits us to connect the capacitor either in series or in parallel with the coil, as required. As an alternative, tapped coils of the type shown in Fig. 206*b* are sometimes used. In cheaper crystal sets continuous frequency coverage is obtained by means of a variometer (Fig. 206*c*). In other circuit arrangements, fixed capacitors are employed and are connected in series with the aerial to shorten the wavelength, or in parallel to lengthen the wavelength.

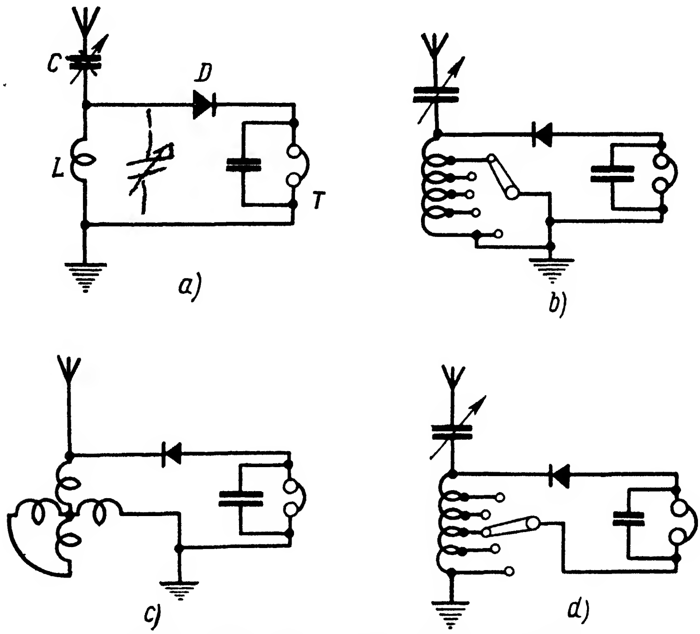

Fig. 206. Simplest crystal detector circuits

The detector circuit possesses a comparatively low impedance and, therefore, exerts a pronounced shunting effect upon the tuned circuit, lowering its Q. It is sometimes possible to increase the loudness of reproduction of a crystal receiver, and at the same time to improve its selectivity, by loosening the coupling between the detector circuit and the tuned circuit. This can be done by connecting the detector circuit to a part of the tuned circuit coil. The advantages thus attained have justified the design of certain crystal receivers in which variable coupling is incorporated (Fig. 206*d*).

A crystal detector receiver is noted for an inherently low sensitivity and poor selectivity. The chief disadvantage of all crystal receivers is the low level of sound reproduction, which is natural because such receivers operate solely by virtue of the energy supplied to them by the insignificantly small power of radio signals picked up by the aerial. Even when such a receiver is located close

to a broadcasting station, the receiver will not produce a sufficiently high output to drive even a small loudspeaker, altrough it will provide a good earphone reception. The earphones will reliably reproduce the signal even when the crystal set is located as far as 500 kilometres from the transmitting station, provided the station is of the high-power variety. When the distance exceeds approximately 500 kilometres, the reception is still possible, but will not be steady and will be very weak.

When it is required to employ a crystal receiver at maximum practical distances from a broadcasting station, the following measures must be taken. The receiving aerial should be long and strung up as high as possible. The high-frequency energy losses should be reduced to the minimum both in the aerial and in the tuned circuit of the receiver. The resistance of the aerial wire, of the earth connection and of the tuned circuit coil should be kept as low as possible. To further reduce the high-frequency losses, only an air-dielectric capacitor should be used in the receiver tuned circuit. And, of course, the quality of the earphones and of the crystal must be the best.

It is a good practice to employ piezo-electric earphones with a crystal receiver, because such earphones are more sensitive than electromagnetic earphones. When piezo-electric earphones are not available, high-impedance electromagnetic earphones may be used in their place. In the latter case, the earphones must be shunted with a 1,000-2,000 pf blocking capacitor, as shown in the drawings of Fig. 206. The high-frequency alternating voltage developed across the tuned circuit is fed to the detector through this blocking capacitor. If no such capacitor were connected, the high-frequency current would have to pass through the earphones, and a considerable part of the high-frequency energy would be lost in the earphone coils. Besides this, the capacitor serves another purpose, as explained below. Current pulses, passing through the detector and reaching the earphones, charge this capacitor. The latter discharges through the earphones during the intervals between the pulses, acting in a way similar to the input capacitor of a smoothing filter, such as is used in a rectifier. The described action of the blocking capacitor, connected across the earphones in a crystal receiver, results in an increase of the audio-frequency voltage appearing across the earphones. This increases the loudness of the reception. It should be noted, however, that the blocking capacitor in a crystal set is not absolutely necessary, because the capacitance existing between the two wires of the earphone cord performs the function of such a capacitor, to a certain extent.

102. STRAIGHT-AMPLIFICATION RECEIVERS

Valve receivers are incomparably more sensitive and selective than the crystal receivers. The signals of extremely remote radio stations, picked up by the aerial of a valve receiver, are easily amplified and reproduced by it at the loudspeaker volume. The principal feature of a valve radio receiver is that the receiver employs some type of anode power supply (for instance a battery or a rectifier) and electron valves to provide a very high amplification (up to millions of times), the energy of the signals picked up by the aerial serving only to control the energy drawn by the valves from the anode power supply.

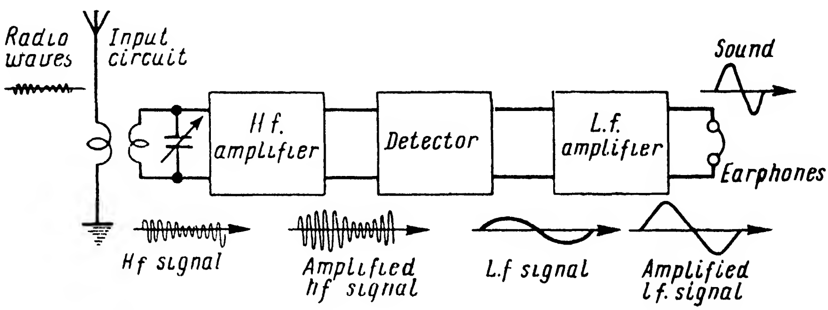

Fig. 207. The block diagram of a straight-amplification receiver

The general design principle of a straight-amplification valve receiver is shown in the *block diagram* given in Fig. 207.

The high-frequency oscillations picked up by the aerial are first fed to the tuned input circuit of the receiver, this circuit being coupled to the aerial. The block diagram shows an inductive variety of such coupling, which is, incidentally, quite frequently used. The input circuit, tuned to the frequency of the incoming signal, provides a certain amount of amplification and also gives some preliminary selectivity, separating the signal of a desired radio station from the signals of numerous other stations simultaneously picked up by the receiving aerial. The high-frequency oscillatory voltage built up across the tuned circuit at the input of the receiver is then applied to the first stage of high-frequency amplification, usually referred to as the h.f. amplifier. The h.f. amplifier usually consists of not more than two stages. Employing electron valves and resonant tuned circuits, this amplifier provides a considerable amplification and an improvement of selectivity.

Having passed through the h.f. amplifier, the amplified high-frequency signal reaches the detector stage. A valve detector, used in this stage, rectifies the signal and usually also gives a certain amount of additional amplification.

The low-frequency (audio-frequency) signal developed at the output of the detector is, next, amplified in low-frequency amplifier stages, usually called for short l.f. amplifiers. There are usually not more than two l.f. amplifiers in a radio receiver. The last l.f. amplifier feeds the amplified audio-frequency signal to a loudspeaker or to a pair of earphones.

It is a customary radio engineering practice to briefly designate the type of a straight-amplification receiver by means of the following abbreviations:

letter V indicates that the receiver has a detector-stage somewhere in its circuit;

number, standing before letter V, shows the number of h.f. amplifiers;

number, standing after letter V, shows the number of l.f. amplifiers.

These abbreviations permit us to coin simple formulas, a glance at which immediately shows the make-up of any straight-amplification receiver. This is illustrated by the following examples:

1-V-1 indicates that the receiver employs one h.f. amplifier, followed by a detector, the latter being followed by a single l.f. amplifier;

0-V-0 stands for a single-valve radio receiver employing neither h.f. nor l.f. amplifiers;

0-V-2 shows that the receiver employs no h.f. amplifiers, and its detector stage is followed by two l.f. amplifiers.

All stages of a radio receiver are supplied with power from filament and anode supply sources, which may be dry batteries, storage batteries or rectifiers. Supply sources are not shown in block diagrams, as a rule.

The most important stage of any radio receiver is its detector stage, because no receiver can operate without a detector. Generally speaking, neither h.f. amplifiers nor l.f. amplifiers are absolutely necessary and a radio receiver will operate without them. However, in this case, the operation of the receiver will be poor—it takes h.f. amplifiers to give any receiver high sensitivity and good selectivity and takes l.f. amplifiers to boost up the detected audio-frequency signal to a sufficiently high level.

103. THE DIODE DETECTOR

In a crystal detector circuit the crystal may be replaced by a diode valve.

The circuit will function normally, just as it did before the crystal was removed from it and substituted by the diode. The operation of the diode is stable, although the valve detector will give a somewhat

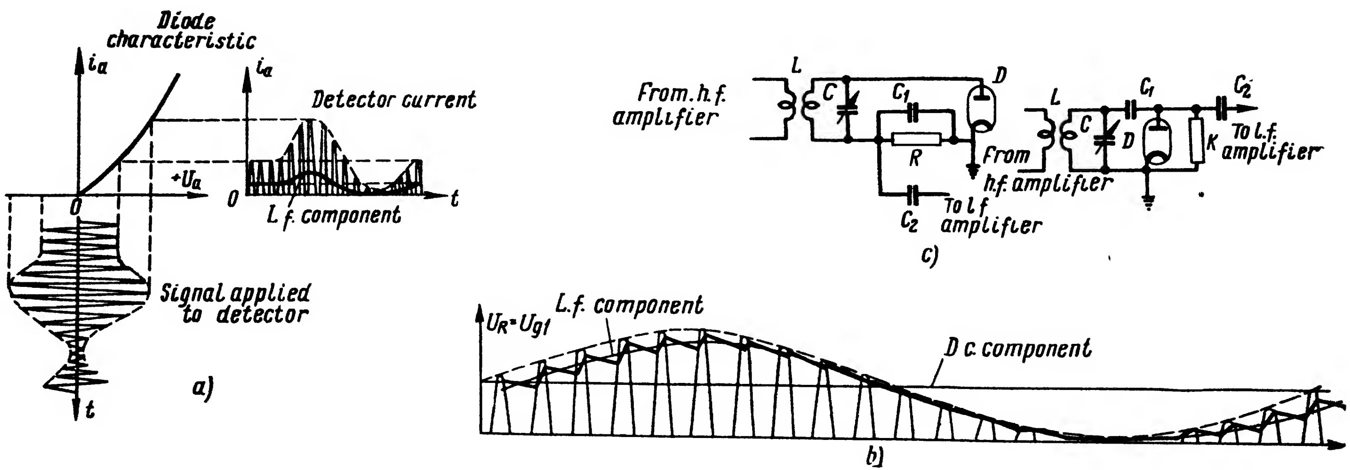

Fig. 208. A graphic representation of diode detection *(a* and *b)*, and the circuit diagrams of the series *(c)* and the parallel *(d)* detectors

decreased sound level, in comparison with a crystal detector, on the reception of weak radio signals.

Diode detectors are very often employed in modern multi-valve superheterodyne radio receivers, where the detector rectifies comparatively strong signals, amplified in the stages preceding the detector. Low distortion of the audio-frequency output signals is the advantage of a diode detector. Its disadvantage is that it does not provide any amplification of the oscillations it detects. Fig. 208*a* gives a graphic representation of the detection process taking place in a diode. The curve of the modulated voltage, fed to the diode from the tuned circuit, is shown along the lower part of the vertical axis. The curve of the pulsating anode current is shown along the right-hand part of the horizontal axis. Apart from the high-frequency component, this current also contains a d.c. component as well as a low-frequency component. It should be noted that, for the sake of simplicity, the curves given in Fig. 208*a* pertain only to the case when no resistance of any kind is connected in series with the diode.

Two circuits of diode detectors are shown in Fig. 208 *c* and *d*. In the circuit given in Fig. 208*c* and called the *series circuit*, the load resistor R is connected in series with the diode. The alternating modulated voltage, taken off from across tuned circuit LC, is fed to the diode, this voltage playing the role of anode voltage. The value of resistance R is of the order of 0.1-0.5 megohm. In order to avoid the loss of a considerable part of the high-frequency voltage, this resistance is always shunted by capacitor C_1 (100-200 pf), which presents but an insignificant capacitive reactance to the high-frequency currents.

Because of the one-way conductance of the diode, the pulsating current flows in the following way. The high-frequency component of this current passes through capacitor C_1 and tuned circuit LC. The d.c. component and the low-frequency component pass through the tuned circuit coil L and through resistor R, building up across this resistor a certain voltage which pulsates at the audio frequency.

Load resistor R is included into the circuit for the specific purpose of obtaining low-frequency voltage, as a result of detector action. This voltage is usually fed through blocking capacitor C_2 to a low-frequency amplifier. The blocking capacitor prevents the d.c. voltage from reaching the l.f. amplifier — (this voltage is also built up across resistor R). The capacitance value of capacitor C_2 must be sufficient (at least several thousand picofarads) to pass the low-frequency oscillations freely.

Capacitor C_1, which shunts load resistor R, serves not only to pass the alternating voltage from the tuned circuit to the diode but also to smooth the high-frequency pulsations of the voltage across resistor R. In other words, this capacitor increases the audio-frequency voltage and acts in a similar way to the input capacitor of the smoothing filter in a rectifier (see Chapter V, Sec. 54).

Owing to the action of capacitor C_1, both the d.c. voltage and the low-frequency voltage, appearing across resistor R, are increased.

Fig. 208*b* gives a graphic representation of the voltage built up across resistor R during the detection of modulated oscillations. Every current pulse, passing through the diode, charges up the shunting capacitor C_1, the said capacitor discharging through resistor R during the intervals between the pulses. As a result of this, the high-frequency pulsations of the voltage are sharply reduced, while the d.c. component of the voltage and the low-frequency component (see the graphic representation) become considerably greater (by 2.5-3 times) in comparison with the case when the shunting capacitor is not used. It should be noted that the process shown in Fig. 208*b* takes place in all cases of detection when the capacitor is present in the circuit.

The circuit shown in Fig 208*d*. and called the parallel circuit employs a parallel connection of the diode and load resistor R. Here, the alternating voltage is fed from tuned circuit LC to the diode through capacitor C_1 (100-200 pf). The high-frequency component of the anode current of the diode passes through this capacitor and the tuned circuit. At the same time, the d.c. component and the low-frequency component flow through load resistor R, being kept away from the tuned circuit by capacitor C_1 which does not pass the direct current and offers a very high capacitive reactance to the low-frequency current. D.c. voltage and audio-frequency voltage are built up across resistor R, the latter voltage being fed to a l.f. amplifier through capacitor C_2.

The diode has a comparatively low internal resistance. This resistance shunts tuned circuit LC, introducing a considerable damping into this circuit. As a result, the selectivity of the tuned circuit is considerably impaired.

This type of detector needs no anode supply and consumes the heater power only.

104. GRID DETECTOR

Straight-amplification receivers usually employ grid detectors. In this type of detector, the process of detection takes place in the control grid circuit. This process is, to a certain degree, similar to the detection process in the diode detector; in the grid detector the role of the diode is played by the space between the grid and cathode, the grid acting as the anode (Fig. 209). Just as it was in the diode detector circuit, here, too, a high resistance (R_g) and a capacitor (C_g) are connected into the input circuit of the detector (in this case — into the grid circuit). In the circuit shown in Fig. 209*a*, the grid resistor is connected in series with the grid-cathode section

and is shunted by the grid capacitor. This circuit is similar to the series circuit of the diode detector (Fig. 208*c*). In the circuit shown in Fig. 209*b*, grid resistor R_g is connected in parallel with the grid-cathode section, this case being similar to the parallel circuit of diode detection (Fig. 208*d*). The capacitance of grid capacitor C_g does not exceed 100-200 pf, while the resistance value of grid resistor R_g may be anything between one to several megohms.

As a result of the detection of the modulated oscillations, a pulsating current appears in the grid circuit, this current consisting of three components. The high-frequency component passes through

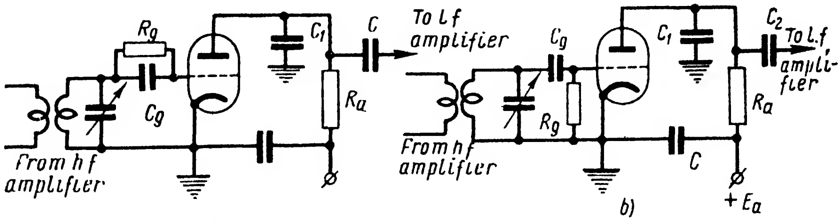

Fig. 209. Grid detection circuit diagrams

the grid capacitor, while the other two components flow through the grid resistor, setting up a voltage across it, this voltage varying at the audio frequency.

The given voltage acts upon the anode current of the valve, introducing audio-frequency pulsations into it. In other words, the alternating audio-frequency voltage, built up across resistor R_g as a result of the detection, is amplified by the triode along the linear part of the anode current characteristic. Simultaneously, the triode amplifies the alternating high-frequency voltage, because this voltage is also present at the grid. As a matter of fact, the following three processes take place in the circuit of the grid detector: diode detection in the grid circuit, amplification of low-frequency oscillations and amplification of high-frequency oscillations.

The optimum operating condition of the grid detector is secured when the operating point is located on the linear part of the anode current characteristic and at the bend of the grid current characteristic.

In the circuits shown in Fig. 209, load resistor R_a is connected into the anode circuit of the detector valve. The amplified audio-frequency voltage, built up across this resistor, is fed to a l.f. amplifier through blocking capacitor C_2 for the purpose of further amplification. If the radio receiver has no l.f. amplifier, a pair of earphones is connected in place of resistor R_a to reproduce the signal detected by the described circuit. The circuits shown in Fig. 209 do not make any use of the amplified high-frequency oscillations. Because

of this, the high-frequency component of the anode current is by-passed by capacitor C_1 and is not allowed to flow through R_a. The capacitance of this capacitor, connected between the anode and cathode, does not exceed a few hundred picofarads.

In some valves, for instance in those employing bariated filament-cathode, the grid current flow does not begin at zero grid potential but rather when the grid is slightly positive (several fractions of one volt). When dealing with such valves, it is a good practice to connect grid resistor R_g to the positive terminal of the filament battery, so that the grid will have a certain positive charge and the operating point is shifted into the region where the characteristic of the grid current is bent.

On the other hand, in valves with oxide-coated cathodes the grid current begins to flow when the grid is slightly negative. In such valves, grid resistor R_g should be connected to the negative terminal of the filament, or, if the valve is of the indirectly-heated type, simply to the cathode.

The grid detector has a high weak-signal sensitivity and gives a greater audio-frequency output than the diode detector. It, however, has the following disadvantage. When a strong signal is applied to its input circuit (i.e., across grid resistor R_g), a high value of bias voltage is developed in the detector stage. This shifts the operating point to the left along the anode current characteristic towards its lower bend, resulting in an amplification with strong non-linear distortion.

Pentode valves can be successfully used as grid detectors.

105. THE ANODE DETECTOR

The process of anode detection is performed as follows. The operating point on the grid characteristic is set (by proper grid bias adjustment) at the lower bend of anode current characteristic (Fig. 210). Under this condition, the negative half-waves are amplified to a much lesser degree than the positive half-waves. The pulsations of the anode current become asymmetrical and a low-frequency component appears in the anode current.

In the circuit of the anode detector, shown in Fig. 210, the grid bias voltage is supplied by grid battery B_b. However, a self-biasing arrangement may be employed in this type of detector (see Chapter VII, sec. 75). The purpose served by components R_a, C_2 ans C_1 is the same as that in the circuit of the grid detector.

Anode detection is not quite applicable to the detection of weak signals, because, when the amplitude of the incoming signals is low, only an insignificant part of the characteristic is used, and the slope of this part is small near the lower bend. In general, the anode detector is less sensitive than the grid detector as it operates only over the lower bend of the valve characteristic, while the grid detect-

or operates over the linear part of the anode current characteristic where the slope has the greatest value. Anode detection should be used on strong signals, the positive half-waves of which reach the linear part of the characteristic.

Anode detection, because of its low sensitivity, is not employed in straight-amplification receivers. The anode detector operates

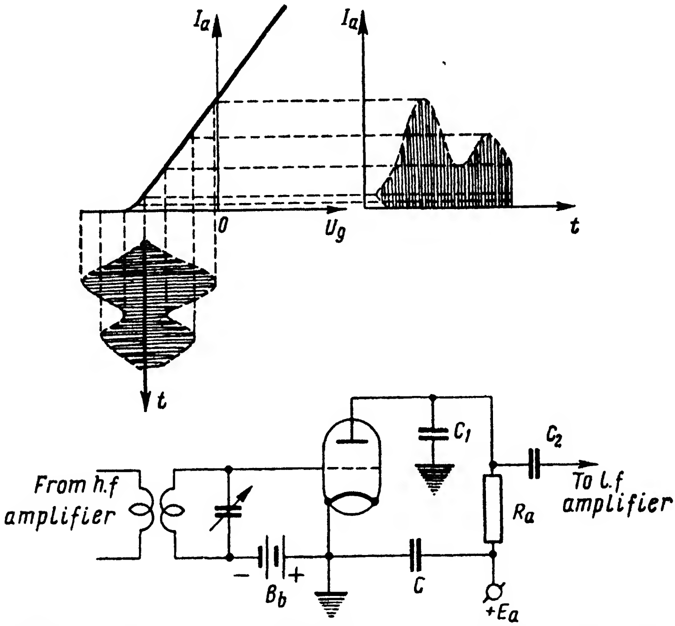

Fig. 210. Graphic representation of anode detection and the circuit diagram of the anode detector

without any grid current, and this is the advantage of this detector. Because of the absence of grid current, the grid-cathode section has a very high impedance and, therefore, exerts but an insignificant shunting effect upon the tuned circuit. This accounts for the better selectivity of the anode-detector, as compared with the diode detector and the grid detector.

Making a general comparison of the three types of valve detectors discussed above, the following conclusions are in order:

1) the grid detector is the most sensitive of the three detectors, i.e., it gives the highest amplification. However, the grid detector introduces considerable distortion on strong signals and somewhat impairs the selectivity of the tuned input circuit;

2) the anode detector gives lower amplification than the grid detector, particularly on weak signals. However, it provides a good detection of strong signals and does not impair the selectivity of the tuned circuit;

3) the diode detector has a low sensitivity, because a diode valve does not amplify signals, and also impairs the selectivity of the tuned circuit. However, this type of detector gives the least amount of distortion, even on strong signals.

106. THE PRINCIPLE AND THE PECULIARITIES OF SUPERHETERODYNE RECEPTION

The superheterodyne reception is based on the conversion of the received high-frequency signal into an intermediate-frequency signal, the latter possessing a constant frequency value. The intermediate frequency (usually referred to simply as i.f.) is also a radio-frequency, although it is usually lower than the frequency of the incoming h.f. carrier. Since a superheterodyne receiver is so designed that the frequency passed by its i.f. stages remains unchanged under all conditions, the parameters of these stages are selected to give an optimum amplification (sensitivity) and selectivity on this particular frequency. Superheterodyne receivers employ a large number of amplification stages and tuned circuits, which accounts for the higher sensitivity and selectivity of these receivers, as compared to receivers of the straight-amplification type. Besides this, the sensitivity and selectivity in a superheterodyne receiver are more constant than in a straight-amplification receiver.

In order to convert the incoming high-frequency signal into intermediate-frequency signal, the superheterodyne receiver generates its own high-frequency signal, which beats with the incoming signal. The circuit of a superheterodyne receiver is so arranged, that the frequency of the locally-generated h.f. signal is always different from that of the incoming h.f. signal by a constant value, which is the intermediate frequency. The two h.f. frequency signals go through the beating process in a special stage employing a valve known as the first detector. It is this stage that produces the intermediate frequency, feeding it to the subsequent stages for the purpose of amplification. The intermediate frequency in superheterodyne receivers is usually selected in one of the following ranges: 110-130 kc, 450-470 kc or 550-570 kc, although some receivers employ intermediate frequencies in a higher range of 1,100-1,600 kc. If, for instance, the frequency of the incoming signal is equal to 2,000 kc and the intermediate frequency is selected to be 460 kc, the local h.f. oscillator in the receiver must generate a signal whose frequency will differ by 460 kc from the frequency of 2,000 kc. Such a frequency may be either 1,540 kc or 2,460 kc. Beating one of these frequencies with the frequency of the incoming signal (2,000) will produce the same intermediate frequency of 460 kc.

The frequency of oscillations, locally-generated in a superheterodyne receiver — be it designed for the reception of long, short or medium

waves — is usually made higher than the frequency to the incoming signals.

Fig. 211 gives the detail block diagram of a superheterodyne receiver. Modulated oscillations having a frequency of, for example, 2,000 kc, are picked up by the aerial and induced in tuned circuit L_1C_1. The oscillations are then fed from this tuned circuit to a h.f. amplifier, the latter usually consisting of one stage of amplification. This amplifier is frequently referred to as the *preselector*, because it performs the preliminary *selection* (i.e., the separation from unwanted signals) of the desired radio signal. In simpler versions of superheterodyne receivers, the preselector stage is frequently omitted from the circuit. The desired h.f. signal, after its separation from other h.f. signals in the preselector stage (or after its separation from such signals in the tuned input circuit L_1C_1 — when dealing with the simpler types of receivers), is passed on to the next stage, called the *frequency converter*. The frequency converter stage actually consists of two stages, the first of which is known as the *mixer* (or the *1st detector*), while the second one is called the *heterodyne* or simply the *local oscillator*.

The heterodyne stage is a low-frequency oscillator, and generates a local high-frequency signal, the frequency of this signal, in the given example, being equal to 2,460 kc. This signal is applied to the mixer stage together with the incoming signal, where the two h.f. signals go through the process of beating. The resulting beats, after their detection in the same stage (i.e., in the 1st detector), acquire a frequency of 460 kc and represent the intermediate frequency oscillations, the latter modulated in the same way as the incoming h.f. carrier signal.*

Simultaneously with the conversion of the received signal, the mixer stage usually provides a certain amount of amplification of the resultant i.f. signal. The i.f. signal is then passed on to a special intermediate-frequency amplifier for the purpose of further amplification of this frequency. The i.f. amplifier usually consists of one or two amplification stages, which are, in effect, high-frequency amplifiers permanently tuned to the frequency of the i. f. signal. Upon leaving the i.f. amplifier, the oscillations of the intermediate frequency reach the *2nd detector*, which converts them into audio-frequency signal, the latter subsequently amplified by one or two stages of low-frequency amplification constituting an l.f. amplifier.

The tuned circuits of the i.f. amplifier of the superheterodyne receiver are permanently tuned to the i.f. frequency by means of trimming capacitors or by paramagnetic cores. The tuned input

* As a matter of fact, several signals of various frequencies are generated in the 1st detector stage. However, since the tuned circuit of this stage is adjusted to a definite frequency, only the signal of this frequency (i.e., the required i.f. signal) is passed forward by the detector stage.

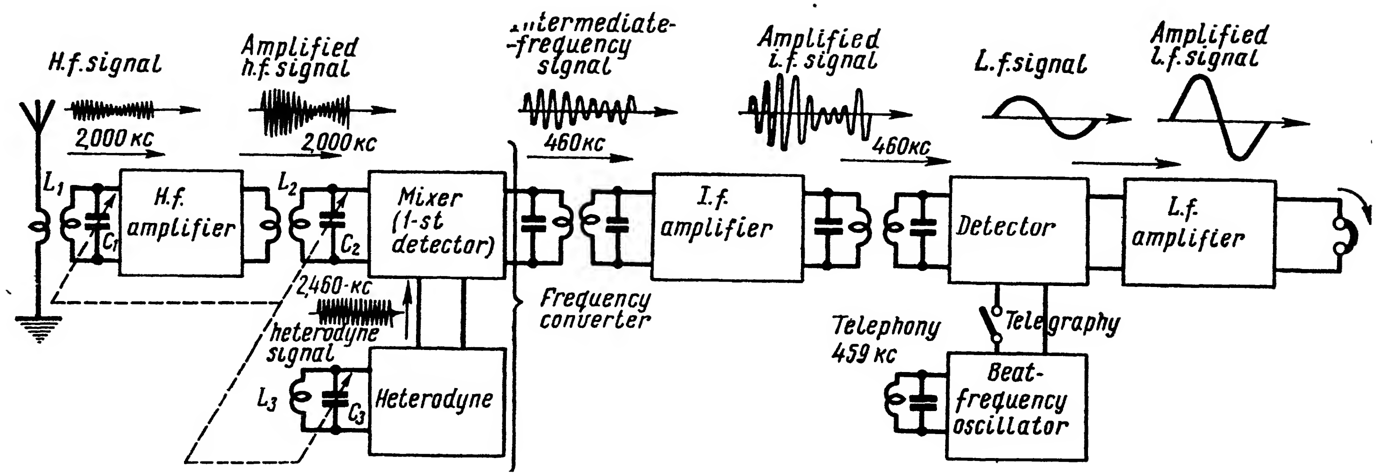

Fig. 211. The block diagram of a superheterodyne radio receiver

circuit L_1C_1, tuned circuit L_2C_2 of the 1st detector and tuned circuit L_3C_3 of the heterodyne are adjusted by means of variable capacitors, which are usually ganged and operated by a single common shaft.

Fig. 212 illustrates the processes taking place in the superheterodyne receiver on the reception of modulated signals. An example of

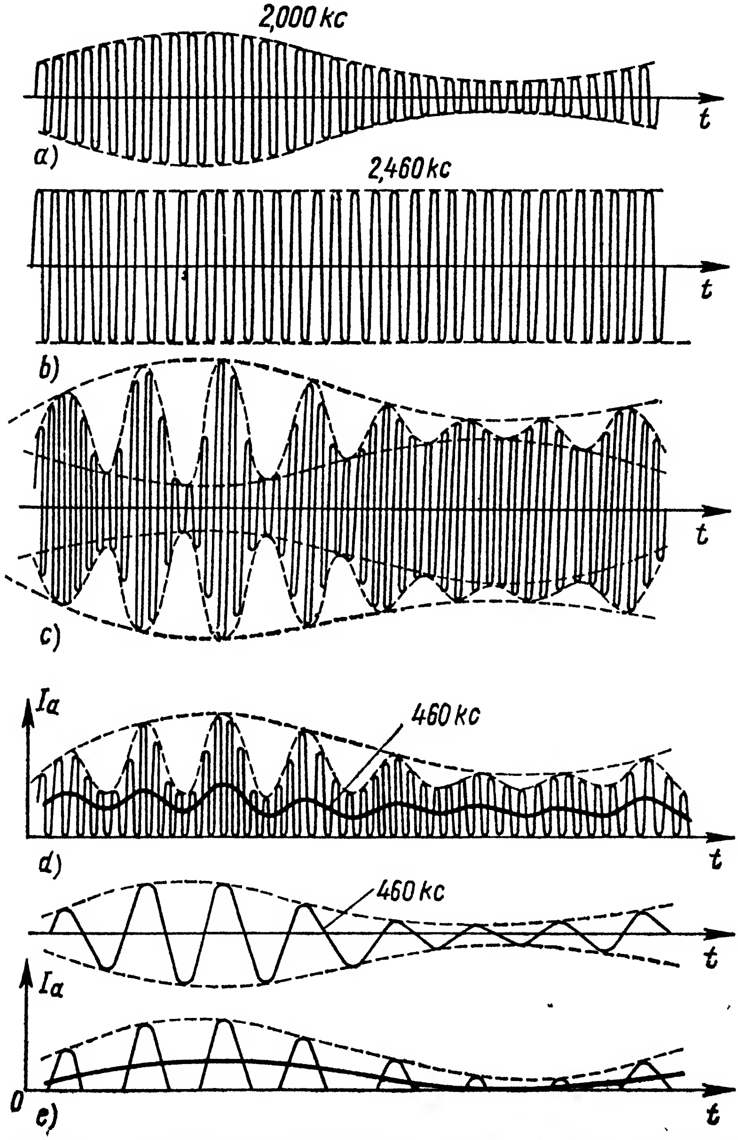

Fig. 212. The graphic representation of processes taking place in a superheterodyne receiver

such signal is given in Fig. 212*a*. The continuous-oscillation curve, shown in Fig. 212*b*, represents the high-frequency voltage developed by the heterodyne. The beats resulting from mixing of this voltage with the modulated high-frequency voltage of the incoming signal are given in Fig. 212*c*. As a result of the detection of such beats (the example gives a case of anode detection), the anode current of the first detector will have a shape shown by the curve of Fig. 212*d*. The thick line drawn under the peaks of this curve shows the component pulsating at the intermediate frequency. This component, after the i.f. amplification, is represented by the curve of Fig. 212*e*. Finally, the curve of Fig. 212*f* gives the current of the 2nd detector. Here, the thick line shows the component pulsating at the audio frequency (i.e., at the frequency of modulation). In the given example we assume that the 2nd detector is of the diode type which is usually the case in superheterodyne receivers.

Unmodulated telegraph signals, usually called *continuous wave signals or simply c.w. signals*, cannot be reproduced by the superheterodyne receiver circuit shown above. The reason for this is the absence of the audio-frequency component at the output of the 2nd detector, when the receiver is tuned to a c.w. signal. This can be shown to be true by plotting curves for such a case the same way as the curves of Fig. 212 were plotted. Accordingly, a superheterodyne receiver designed for the reception of c.w. signals must employ a second heterodyne, known as the *beat-frequency oscillator*. This heterodyne is nothing but a simple oscillator generating a frequency which is constant and differs by about 1,000 cps from the intermediate frequency of the given receiver. Taking the case given above, when the i.f. frequency is equal to 460 kc, the frequency of the beat-oscillator must be either 459 or 461 kc. Such a signal, developed by the beat oscillator, is fed to the 2nd detector, where it is mixed with the i.f. signal. The mixing results in beats, which, after detection, produce a 1,000-cps output signal—an audio frequency.

The pitch of such audio-frequency signal may be varied by slightly changing the high-frequency tuning of the receiver. This is easily understood from the following. When the tuning is changed (by rotating the shaft of the ganged capacitors), the frequency of the signal generated by the 1st heterodyne will vary. The frequency of the incoming signal picked up by the receiver remains unchanged, because this frequency depends not only on the adjustment of the receiver circuits but also on the operating wavelength of the transmitting radio station. Consequently, when the tuning of the ganged capacitor is slightly changed, the difference between the frequency of the incoming signal and the frequency of the heterodyne will vary, thus varying the i.f. frequency. But the frequency of the beat oscillator is constant and, therefore, a change of the i.f. frequency, beating with the frequency of the beat oscillator, will cause a change of the resultant beat frequency at the output of the 2nd

detector. Thus, on the reception of c.w. signals, *the pitch of the audio* signal may be varied by the operator, who *will slightly detune* the ganged capacitor from the incoming carrier frequency. The operator has the choice of setting the tuning to such a pitch of the output audio-frequency signal, where the signal best penetrates the signals of interfering radio stations and noise. In some superheterodyne receivers the pitch is varied by changing the frequency of the beat oscillator, and in this case the high-frequency circuits of the receiver are kept in exact resonance with the frequency of the incoming carrier wave.

The smaller the difference between the intermediate frequency and the frequency of the beat oscillator, the lower will be the pitch of the beat tone. When these two frequencies are made equal, the beats disappear altogether. Such a case is known as the *zero-beat condition*. Zero beats play an important role in radio engineering. In a radio receiver, the zero-beat condition is an indication that the set is tuned exactly to the frequency of the radio station being received. When the receiver is detuned to either side of the zero-beat position, the audible beat tone appears and its pitch grows higher as the detuning is made larger. On further detuning, the audible beat tone changes into a very high-pitched whistle and finally the signal becomes no longer audible because the ear does not react to such a high frequency of beats.

When the superheterodyne receiver is tuned to a radio station radiating modulated waves (radio telephony or modulated radio telegraph signals, the latter known as m.c.w.), the beat oscillator circuit is switched off to prevent the beat notes from interfering with the received intelligence. A block diagram of beat oscillator connection to the 2nd detector is shown in Fig. 211.

Superheterodyne receivers designed just for the reception of broadcast programmes do not employ beat-frequency oscillators and can reproduce only the signals of radio-telephone and m.c.w. telegraph stations. On the other hand, recivers designed for the reception of c.w. telegraph stations are always provided with beat oscillators. The latter type of receivers also has a change-over switch marked "Telephony-Telegraphy", this switch cutting in the beat-frequency oscillator on the reception of c.w. signals (the "telegraphy" position of the switch), or disconnecting the oscillator on the reception of modulated signals (the "telephone" position).

An alternative method of c.w. signal reception also exists. In this method, the incoming c.w. signals are subjected to modulation right in the receiver circuit. Such modulation is performed by a 1,000-cps audio oscillator incorporated into one of the i.f. amplifier stages. The audio signal developed by the oscillator modulates the i.f. signal, the latter acting as a "carrier". As a result, the i.f. signal carries the 1,000-cps modulation by the time it reaches the 2nd detector. Receivers of this type need no beat oscillators for the reception of c.w. telegraph stations, because the receiver circuit itself modulates all c.w. carriers by a pleasant and piercing 1,000-cps tone. Such receivers are also provided with the "Tele-

phony-Telegraphy" switch, the latter stopping the 1,000-cps oscillator when thrown to the "Telephony" position. These receivers are noted for the following peculiarity. When the high-frequency tuning circuits of such a receiver are slightly detuned from the frequency of a radio station being received, or when the carrier frequency of the station is slightly shifted, the telegraph signals of the station are still reproduced as 1,000-cps tones at the output of the receiver, because the local audio-frequency oscillator incorporated in the receiver is always operating on this frequency. By the same token, the signals of various radio stations are invariably reproduced by the receiver of this type as 1,000-cps tones. The detuning of the receiver changes only the strength of the output signals but not their pitch. This accounts for the high reception stability, provided by the receivers of this type.

When compared to a straight-amplification radio receiver, a superheterodyne receiver offers the following advantages:

1) high sensitivity, attributed to the greater number of stages and to the higher gain attained in the i.f. amplifier;

2) high selectivity, attributed to the greater number of tuned circuits;

3) better constancy of sensitivity and selectivity over the entire tuning range of the receiver;

4) adaptability to the incorporation of various improvements, such as the automatic gain control, electronic optical tuning-indicators, and other improvements which can be introduced only when the receiver has a high gain.

At the same time, the superheterodyne receiver is noted for several peculiar disadvantages, which are here discussed in detail.

Receiver noise. All types of radio receivers develop a certain amount of internal noise, which is reproduced as interference at the output of the receiver. This internal noise, referred to as the receiver noise, is caused by various irregularities of the emission in the electron valves and also by the haphazard thermal movement of the electrons in receiver wiring and resistances. Because of the greater number of valves employed by superheterodyne receivers and because of the greater gain provided by such receivers, the internal noise is much higher in a superheterodyne receiver than in a straight-amplification set.

The images. In a superheterodyne receiver, heterodyne frequency f_h is usually higher than frequency f_s of the incoming signal by the value of the intermediate frequency (f_{if}). This is the frequency relation existing between the tuned circuit of the heterodyne and the other h.f. tuned circuits, the latter circuits being adjusted to the frequency of the incoming signal. If $f_s = 2{,}000$ kc and $f_{if} = 460$ kc, then f_h must be made equal to 2,460 kc. Now, let us assume that the aerial of a superheterodyne set receives simultaneously the 2,000-kc signal of the desired station and the signal of an interfering radio station operating at a frequency of f_i, this frequency being equal to 2,460 kc + 460 kc = 2,920 kc. Obviously, the difference between frequency f_i and frequency f_h constitutes 460 kc. This frequency difference is the same as the beat frequency resulting from mixing

frequencies f_i and f_h in the 1st detector of the receiver. After the detection of these beats by the detector valve, an intermediate frequency of 460 kc will be obtained at the output of the detector. This frequency will be amplified by the i.f. amplifier and, as a result, the interfering radio station, working on 2,920 kc, will be reproduced at the output of the receiver. Such an interference is called *the image interference*, or simply *the image*. The reason for the name is that the frequency of the interfering station is removed from the frequency of the heterodyne by the same number of kilocycles as the frequency of the desired station. In other words, image frequency f_i differs from frequency f_s of the desired station by double the value of the intermediate frequency f_{if}.

It should be noted that the image becomes noticeable only when the signal level of the interfering carrier f_i is high. The high-frequency tuned circuits of the receiver are adjusted to the desired frequency f_s and are, therefore, detuned by $2f_{if}$ (in the given example — by 920 kc), as far as the image frequency f_i is concerned. This detuning of the high-frequency tuned circuits serves to attenuate the level of the offending carrier reaching the radio receiver. If the intermediate frequency is made lower (for instance, made equal to 120 kc), the degree of the detuning will be less (240 kc, in the given case), and the image will be more pronounced, particularly on short waves. This is one of the reasons why the intermediate frequency of a superheterodyne receiver should not be too low.

Apart from employing a sufficiently high intermediate frequency, another way of suppressing the images is that of using high-frequency selection (called the preselection) in a superheterodyne receiver. Such preselection is provided by including into the receiver circuit at least one stage of high-frequency amplification, preceding the 1st detector. Still another way of suppressing the images is that of incorporating a band filter at the input of the receiver.

Besides the question of images, the described peculiarity of the superheterodyne receivers also accounts for the so-called "double-tuning" effect in such receivers; a superheterodyne receiver may be tuned to one and the same radio station at two different settings of the receiver tuning dial. Here, the basic dial setting is the one at which the high-frequency tuned circuits of the h.f. amplifier and of the 1st detector are tuned to the signal of the desired station. In this case, if $f_s = 5{,}000$ kc and $f_{if} = 460$ kc, the heterodyne frequency f_h, corresponding to the basic tuning (the basic dial setting), is given by $f_h = 5{,}000 \text{ kc} + 460 \text{ kc} = 5{,}460$ kc. The second tuning of the heterodyne, when the receiver will reproduce the same station, corresponds to the following frequency of the heterodyne: $f_h = 5{,}000 \text{ kc} - 460 \text{ kc} = 4{,}540$ kc. The difference between this frequency and the frequency of the incoming signal is also equal to 460 kc, which creates a possibility of receiving the required radio station at this new setting of the heterodyne tuning control.

Such double-tuning effect in a radio receiver is, of course, undesirable because each radio station should be tuned in only at one definite point of the dial. This effect is, accordingly, counteracted by ganging the tuning capacitors of the high-frequency amplifier stages and the tuning capacitor of the 1st detector stage with the tuning capacitor of the heterodyne. Such ganging provides the so-called "single-shaft control". When the single-shaft control is employed in a superheterodyne receiver (which is always the case), the double-tuning effect is neutralised as follows. Taking the case given above and assuming that the h.f. amplifier and the 1st detector are adjusted to the second (the undesirable) point of reception, we see that these stages are tuned to a frequency of 4,080 kc, i.e., detuned by 920 kc in respect to the required incoming signal. Such detuning greatly attenuates the second-point reception. As a matter of fact, when due precautions are taken, the required radio station can be heard on the second point of dial setting only when the station is of a very high power variety and particularly if it is a short-wave station.

Superheterodyne whistles. The frequency conversion taking place in a superheterodyne receiver is the cause of interfering whistles. Such whistles are generated as a result of beating between the following types of oscillations:

a) harmonics of the incoming signal with the fundamental frequency of the heterodyne;

b) heterodyne harmonics with the fundamental frequency of the signal;

c) heterodyne harmonics with the harmonics of the signal.

If the frequency of such undesirable beats is close to the intermediate frequency, audio beats may be generated in the receiver and reproduced as whistles at the receiver output.

Let us illustrate this effect by the following example. Assume that $f_{if} = 460$ kc, $f_s = 922$ kc and $f_h = 922 + 460$ kc $= 1{,}382$ kc. The second harmonic of the incoming signal is equal to $2f_s$, i.e., equal to $2 \times 922 = 1{,}844$ kc. The difference between this frequency and the frequency of the heterodyne is equal to $1{,}844 - 1{,}382 = 462$ kc. Oscillations of this frequency will be obtained after the conversion. These oscillations will pass through the i.f. amplifier and will beat in the tuned circuit of the 2nd detector with the 460-kc i.f. frequency, thus giving rise to 2-kc audible beats, the latter generating piercing 2-kc whistles in the loudspeaker or earphones connected to such a receiver.

The complexity of superheterodyne circuits, and of their construction and adjustment. One of the disadvantages of the superheterodyne receiver is seen in that this type of receiver is considerably more complex than the straight-amplification receiver, as far as its cir-

cuit and the general design are concerned. A "superhet" (as a superheterodyne receiver is often referred to) is more difficult to align and adjust both during the manufacture and during repairs than the receiver of the other type.

107. THE INPUT CIRCUIT AND HIGH-FREQUENCY AMPLIFICATION

The high-frequency amplifier of a radio receiver is called upon to raise the level of the high-frequency signals, received by it from the aerial, and to increase the selectivity of the receiver.

The signals picked up by a receiving aerial are fed to the input circuit of the h.f. amplifier of a radio receiver. The aerial is usually

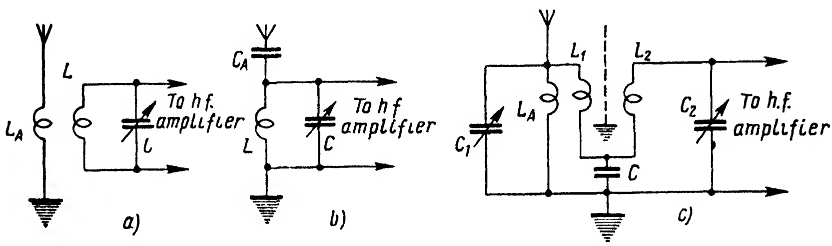

Fig. 213. Methods of coupling the tuned input circuit of a radio receiver with its aerial: — *a)* inductive coupling; *b)* capacitive coupling; *c)* band filter connected into the input circuit

inductively coupled to the input circuit, or else is coupled to it through small capacitance (Fig. 213*a* and *b*). Direct coupling of the aerial to the input circuit is not used as this would include the considerable capacitance of the aerial into the circuit. This capacitance not being constant, such a connection would affect the tuning of the input circuit. In radio receivers, the aerial itself is not tuned, as a rule. In some cases, however, when it is required to obtain a maximum sensitivity and selectivity, the aerial may be tuned to the frequency of an incoming signal by means of a separate capacitor.

Various types of filters may also be used in the input circuit to improve the selectivity of a radio receiver and to suppress interference. For instance, the input circuit may be designed as a ***band filter***, consisting of two closely-coupled tuned circuits. Fig. 213*c* shows such a filter, employing capacitive coupling. A filter designed as a tuned circuit and tuned to the frequency of a powerful nearby radio station is capable of suppressing the interference that such a station would otherwise create to the reception of other stations. An example of such a filter, known as ***band-elimination filter***, is given in Fig. 214*a*. Here, tuned circuit C_fL_f is adjusted to the

frequency of the interfering station, offering a very high impedance to the signals of this frequency and a low impedance to the signals of all other frequencies. An alternative interference-suppressing filter circuit is shown in Fig. 214*b*. In such a circuit, filter elements L_f and C_f are connected in series with the aerial and with each other. A case of series resonance for the offending signal is established when the L_fC_f-aerial combination is tuned to resonance with such a signal. The combination, thus created, presents an insignificant impedance to the signals of the interfering station and virtually short-circuits such signals to earth, preventing them from entering the input circuit of the radio receiver. At the same time, such a tuned combination offers a considerable impedance to the signals of other stations, operating on different frequencies, and directs them to the input circuit of the same receiver.

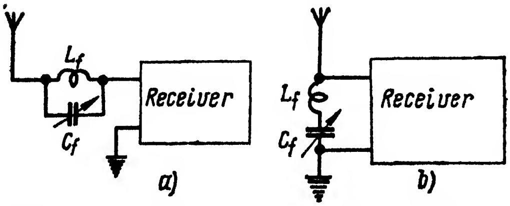

Fig. 214. Interference-suppressing filters at the input of radio receivers

A popular high-frequency amplifier, employing a tuned anode circuit, is shown in Fig. 215*a*. Alternating high-frequency voltage, reaching the grid of the amplifier valve from the input circuit, sets up a high-frequency current in the anode circuit. Tuned anode circuit L_2C_2 acts as the load resistor for the alternating component of the anode current. The higher the impedance of this tuned circuit, the higher will be the gain of the stage. The given tuned circuit is adjusted to the frequency of the incoming signal, and its impedance is naturally great because of the parallel resonance of components L_2 and C_2. This impedance, otherwise referred to as the equivalent resistance of the tuned circuit, is expressed by the following formula:

$$R_e = \frac{L}{rC} \cdot$$

The values of L and C are chosen in accordance with the required wavelength range. The value of R_e should not be raised by increasing the value of L and by decreasing the value of C, because when the capacitance value of the capacitor is too low, the wavelength coverage will be too small. The best way to increase the valve of R_e is to decrease ohmic resistance r of the tuned circuit as much as possible, i.e., to decrease the high-frequency losses in the coil and in the capacitor. The decrease of such losses is also desirable from the point of selectivity improvement.

In practical circuits, the value of R_e cannot be made to exceed a few thousand ohms or, at best, a few tens of thousands. Since the anode resistance of high-

frequency pentodes reaches several hundred thousands and even millions of ohms, the gain of the amplifier stage is comparatively low in comparison with the amplification factor of the valve. Generally, the gain of an h.f. voltage amplifier stage does not exceed some tens, despite the fact that the amplification factor μ of the valve is equal to several hundreds or even several thousands. As an example, a type 6K7 valve has the following parameters: $\mu = 1{,}200$; $R_i = 800{,}000$ ohms; $S = 1.5$ ma/v; if the tuned anode circuit employed by a high-frequency amplifier stage using a 6K7 valve has an equivalent resistance

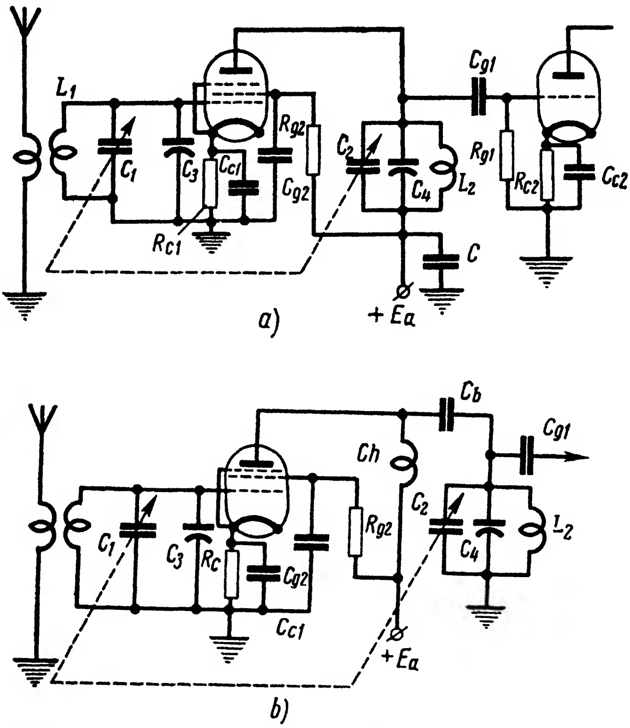

Fig. 215. H.f. amplifiers employing tuned anode circuits: *a)* with series feed; *b)* with parallel feed

$R_e = 10{,}000$ ohms at resonance, the gain of the stage is given by the equation shown in Sec. 72.

$$k \approx SR_e = 1.5 \times 10 = 15.$$

As may be seen, the gain of the stage is equal only to 15, although the μ of the valve is 1,200. Such poor utilisation of the amplification factor of the valve is attributed to the high anode resistance of the valve and to the impossibility of designing a tuned circuit with a sufficiently high value of R_e.

The gain of high-frequency amplifier stages becomes lower on short waves. This is attributed to the fact that on short waves inductance L is considerably reduced, while the capacitance is decreased but a little and can even remain the same as on long waves

(this is observed, for instance, in modern broadcast receivers, in which the same capacitors remain in the circuit while the coils are switched out and in to cover long-wave and short-wave ranges). The resistance of a tuned circuit cannot be likewise considerably reduced, because the h.f. coil losses grow (because of the skin effect) as the frequency is increased, although the resistance of the coil wire on short waves becomes smaller because of the decrease of the number of coil turns. As the net result of all this, the equivalent

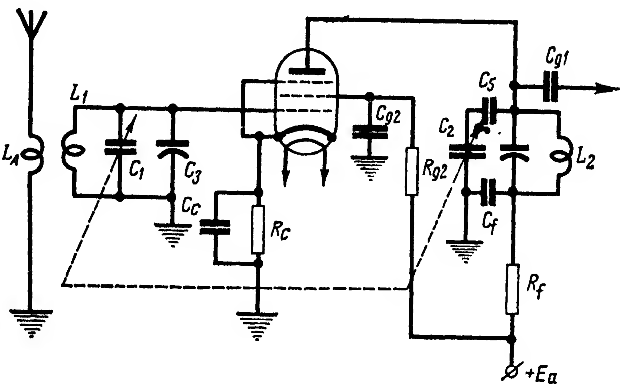

Fig. 216. The circuit of a h.f. amplifier with decoupling filter

resistance R_2 of a tuned circuit becomes smaller on short waves, which accounts for the decrease of the amplification.

For the sake of tuning convenience, variable capacitor C_1 of the tuned input circuit (the grid circuit) and variable capacitor C_2 of the tuned anode circuit are ganged and adjusted by means of a common control, as shown by the broken line in Fig. 215*a*. However, the rotor of capacitor C_2 is placed at a high positive potential and, therefore, cannot be connected through a common metal shaft to the rotor of capacitor C_1, as this would short-circuit the anode voltage source. The circuit shown in Fig. 216 is free from this shortcoming. In this circuit, capacitor C_f, which is used for the purpose of passing the high-frequency component of the anode current, is connected into the tuned anode circuit. This isolates the rotor of tuning capacitor C_2 from the d.c. anode voltage and permits us to connect this rotor to the metal shaft carrying the rotor of capacitor C_1. It is good practice to isolate the stator of capacitor C_2, too, from high d.c. voltage in order to protect the anode voltage source from the short-circuit that might occur upon an accidental contact between the rotor and stator plates of this capacitor. In this case, the necessary isolation is provided by capacitor C_5, the value of which

is similar to the value of C_f and is expressed at least in some thousands of picofarads. Because of such high capacitance value, these two capacitors do not appreciably decrease the total capacitance of the tuned circuit, although the two capacitors are connected in series with each other and with the tuning capacitor C_2. Small trimming capacitors, connected in parallel with capacitors C_1 and C_2, permit us to adjust tuned circuits L_1C_1 and L_2C_2 to exact resonance. In some receivers, the role of the trimming capacitors is performed by special cores which can be moved back and forth within the coils of the tuned circuits, thus varying the inductance of these coils.

Fig. 215*a* shows the coupling of the h.f. amplifier stage to the following stage. The latter stage may be represented by the second stage of h.f. amplification, or by a detector, or else by a frequency converter. The voltage amplified in the first amplifier stage is fed to the grid circuit of the following stage through capacitor C_{g1}, which insulates the grid of the driven stage from the high anode voltage of the driving stage. In such a circuit arrangement, grid resistor R_{g1} cannot be connected in parallel with coupling capacitor C_{g1}, as this would also apply the high anode voltage to the grid of the driven valve; R_{g1} should be, accordingly, connected between the grid and cathode of the valve, as shown in the drawing. Thus, tuned circuit L_2C_2 simultaneously acts as the tuned anode circuit of the driving valve and as the tuned grid circuit of the driven valve.

The screen grid of the valve employed in a h.f. amplifier stage is usually supplied with voltage through dropping resistor R_{g2} or else is supplied from a voltage divider. The screen grid must be in all cases connected to the cathode through capacitor C_{g2}, whose capacitance value should not be less than several thousand picofarads.

High-frequency amplifiers do not use triodes, because such valves can provide only a low value of amplification and, besides, possess a considerable anode-grid interelectrode capacitance C_{ag}, which gives rise to parasitic oscillations. The application of pentode valves in h.f. amplifiers prevents such parasitic oscillations, provided the anode circuits are thoroughly shielded from the grid circuits. Such shielding will eliminate the undesirable capacitive and inductive parasitic couplings, which would otherwise transform the h.f. amplifier into a self-excited oscillator.

The circuit shown in Fig. 215*a* employs series anode feed. The direct anode current flows through coil L_2 of the tuned anode circuit, placing this circuit at a high positive potential. Some high-frequency amplifiers use the alternative parallel anode feed circuit, shown in Fig. 215*b*. In this circuit, the direct component of the anode current flows through choke *Ch*, while the alternating high-frequency component, to which the choke offers a high inductive reactance, passes from the valve to the tuned circuit through blocking

capacitor C_b. This circuit is more rational than the other one because here the tuned anode circuit is not placed at a high voltage and the rotor of the anode tuning capacitor can be installed on the same metal shaft with the rotor of capacitor C_1. However, in the parallel feed circuit choke *Ch* presents something of a problem as it is not easy to design this choke so that it would have a high inductive reactance over a broad range of frequencies. The given choke must consist of a large number of turns (several hundreds or thousands), and the turns must be so wound that their distributed capacitance

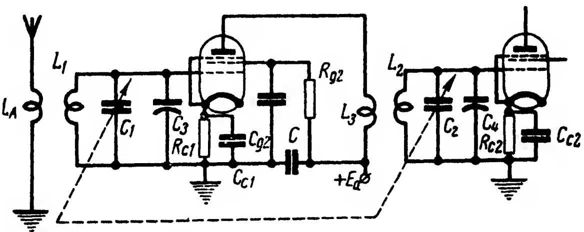

Fig. 217. Transformer-coupled h.f. amplifier

is kept to a low value. Sometimes the choke winding is sectionalised for this purpose. In some cases, a resistor is substituted for this choke. The advantage of such resistor (which has a value of several tens of thousands of ohms) is that its resistance value is approximately constant on all the frequencies of the required range. However, such a resistor produces a considerable drop of the anode voltage, and, besides, shunts the tuned anode circuit, impairing its Q.

Another popular type of h.f. amplifier is shown in Fig. 217 and is known as a *transformer coupled amplifier*. Here, the anode circuit of the h.f. amplifier valve is coupled to the grid circuit of the following stage by means of a h.f. transformer, consisting of coils L_3 and L_2. The secondary coil L_2 is a part of the tuned circuit of the driven stage. The two coils are usually wound side by side on a common coil form. In this amplifier, tuned circuit L_2C_2 is insulated from the high anode voltage and, therefore, the rotor of capacitor C_2 can be ganged with the rotor of capacitor C_1 by a common metal shaft. With correct value of mutual conductance set between coils L_2 and L_3, this circuit provides higher gain and higher selectivity than the circuit employing a tuned anode circuit.

In h.f. amplifier stages the value of bias is usually about two or three volts. Such bias is sufficient to prevent the grid current flow and to reduce the anode current consumption to a sufficiently low value, which is particularly important in battery-operated sets. Automatic biasing circuits used by h.f. amplifiers do not differ from the circuits employed by l.f. amplifiers, discussed under Chapter VII (Secs 75 and 78). For the purpose of elimination of parasitic coupling through the common anode circuits, h.f. amplifier stages

also employ anode decoupling filters of the type studied under Chapter VII (Sec. 78).

An example of a complete h.f. amplifier stage, employing an automatic biasing circuit and anode decoupling filter C_fR_f is shown in Fig. 216. The value of self-biasing resistor R_c is between several hundred and several thousand ohms, while the capacitance of bypass capacitor C_c must be at least 10,000-20,000 picofarads. Resistor R_f has a value of several tens of thousands of ohms, while the value of capacitor C_f is at least several tens of thousands of picofarads.

High-frequency amplifiers introduce distortion into the radio-telephone signals they amplify. As a result of cutting off the sidebands by the sharp selectivity of its tuned circuits, such an amplifier is noted for frequency distortion, caused by the attenuation of the higher frequency components of the modulated wave passed by the amplifier. The deviation of the valve characteristic from the straight-line shape also results in distortion. This is a non-linear distortion, because, in the given case, the shape of the modulated oscillations is changed. The stronger the signal, the more pronounced is such non-linear distortion. When the radio receiver operates from a.c. mains, small 50-cps or 100-cps pulsations of anode and screen-grid voltages cause variations of the parameters of the amplifier valve at such frequencies. As a result, the h.f. signal being received is modulated by these frequencies, the effect known as the *secondary* or *parasitic modulation*. This effect is evidenced by the appearance of a.c. hum and wheezes in the receiver's loudspeaker. Thus, supply voltage pulsations, having a frequency of 50 or 100 cps, are not only capable of being reproduced because of their influence upon l.f. amplifier stages, but can be also reproduced as hum distortion, acting upon h.f. amplifier stages.

If to the grid of a high-frequency amplifier valve, besides the useful signals, is also applied the modulated voltage of an interfering station working on a different frequency, the useful signals can be modulated by the interfering signal as a result of the curvature of the valve characteristic. This causes what is known as *cross distortion*. Such distortion is characterised by a total disappearance or by a considerable reduction of the interference when the useful signals are absent, for instance, as a result of a slight detuning. This effect is attributed to the fact that the cross distortion is produced as a result of the modulating action of the interfering signal upon the grid of the first h.f. amplifier valve. The tuned anode circuit of this valve, being adjusted to the frequency of the useful signal, modulated by the interference, passes the useful signal only. However, the same tuned circuit will not pass the interfering signal by itself, because the frequency of this signal is not the one to which the circuit is adjusted.

108. FREQUENCY CONVERSION

In frequency converters, the heterodyne signal is fed to the mixer, where it is made to beat with the incoming signal picked up by the receiving aerial. In such a circuit arrangement, it is necessary to eliminate the coupling that might exist between the tuned circuit of the heterodyne and the tuned circuit of the h. f. amplifier (or between the tuned input circuit, if no h. f. amplifier is employed by the radio receiver). If such a coupling exists, all changes of tuning of the h.f. amplifier circuit will affect the tuning of the heterodyne circuit, which is always observed between any two tuned and coupled circuits. Such dependence of the heterodyne tuning upon the tuning of the h. f. amplifier is, of course, undesirable and, therefore, appropriate measures should be taken in all superheterodyne sets to eliminate the described type of coupling.

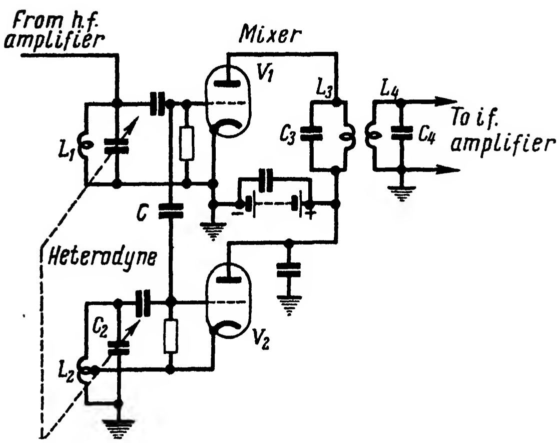

Fig. 218. Single-grid frequency converter

In the simplest single-grid converter, both the incoming signal and the signal of the heterodyne are fed to the same grid of the converter valve. An example of such converter is given in Fig. 218. This circuit employs a mixer, operating as a grid detector and a heterodyne, which is a three-point oscillator with the autotransformer feedback (i. e., a cathode-coupled oscillator). The incoming signal, upon its amplification in a h.f. amplifier, is fed from the tuned anode circuit of the amplifier to the grid of mixer valve V_1. The heterodyne signal is fed to the same grid through a low-capacitance capacitor from the heterodyne stage, in which valve V_2 is used. The two signals go through the beating process in valve V_1, where they are detected. The resulting intermediate frequency is developed in the anode circuit of valve V_1, this circuit tuned to the given frequency by means of capacitor C_3 resonating with coil L_3. The intermediate-frequency signal is then fed to the circuit L_4C_4 and to the grid of the first i.f. amplifier valve.

Such a single-grid converter does not secure the necessary decoupling of the heterodyne tuned circuit from the tuned circuit of the h.f. amplifier. The problem is solved by employing special multi-grid valves for the purpose of frequency conversion. Such

valves are called converter valves or mixer valves, and are classified as hexodes, heptodes, octodes, triode-hexodes and triode-heptodes (see Chapter IV, Sec. 47).

These valves convert frequency by virtue of mixing the incoming signal and the heterodyne signal in the electron stream within the valve, the signals being fed to separate grids of the valve. The presence of two control grids in the same valve envelope is the distinguishing feature of such converter valves, which are, accordingly, called double-control valves.

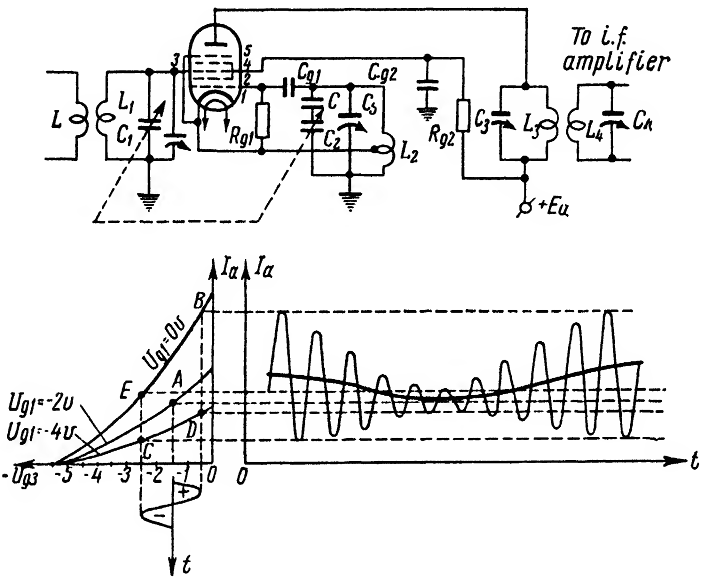

Fig. 219. The frequency converter employing a heptode-converter valve and a graphic representation of the frequency conversion process

The most popular frequency conversion circuit used at present employs a heptode converter type 6A2П or type 6A7(6A10C). This circuit is shown in Fig. 219. In these valves, the heterodyne triode section is represented by the cathode and grids *1* and *2*, where grid *2* acts as the anode of the given triode. The heterodyne employs a cathode-coupled circuit, its tuned circuit being comprised of coil L_2 and capacitors C_2, C and C_5, whose function is explained later. High voltage from the anode power supply is fed to the anode of the heterodyne section (i.e., to grid *2*) through dropping resistor R_{g2}, while the alternating component of the heterodyne anode current passes through capacitor C_{g2}. Thus, grid *1*, usually referred to as *the heterodyne grid*, is placed at the alternating potential of the heterodyne frequency, causing the electron stream inside the valve to pulsate at this frequency.

The function of the second control grid of the converter valve is performed by grid *3*, known as *the signal grid*. To this grid is applied the incoming signal from tuned circuit L_1C_1. Since grid *3* is also a control grid, it follows that the electron stream pulsates also at the frequency of the incoming signal. Hence, the electron stream itself serves as the medium in which the mixing of heterodyne-frequency pulsations and of signal-frequency pulsations takes place.

Grids *2* and *4* are connected to each other within the valve and act as screen grids. Each one of these two grids serves a purpose of its own. Grid *4* is the usual screen grid, whose function is to increase the amplification factor of the valve and to eliminate the parasitic capacitance between the anode and the signal grid. Grid *2* serves to eliminate the capacitance between grid *3* and the heterodyne section, i.e., it eliminates the undesirable coupling between tuned circuit L_1C_1 and the tuned circuit of the heterodyne. Grid *5* is the suppressor grid. Thus, a heptode valve of the described type acts as a pentode. However, this is an unusual kind of pentode, which is provided with two control grids, these grids separated from each other by an auxiliary screen grid. As may be seen, grid *2* simultaneously acts as the anode of the triode section of the valve and also as a screen grid. Because of this, the heterodyne part of such a valve must in all cases employ a circuit in which the anode is earthed in respect to the high frequency. Valves 6A7 and 6A10C are single-ended valves. This is the point of difference between these valves and the earlier converter valves, in which the signal grid was brought out at the top of the valve envelope. Besides this point, a number of design improvements were incorporated in the new converter valves, providing for better efficiency of these valves under conditions of short-wave operation. In the bantam series of valves, designed for dry-battery operation, valves 1A1П and 1A2П are similar to the 6A7 and 6A10C valves described above.

The beats produced by the mixing of the two kinds of electron stream pulsations are detected in the converter valve and, as a result, the i.f. component appears in the anode current of the valve—tuned circuit L_3C_3 being adjusted to this component. The given tuned circuit offers a high impedance to the i.f. component; and, because of this, the converter stage simultaneously amplifies the i.f. signal. At the same time, tuned circuit L_3C_3 offers but a low value of impedance to the incoming-signal frequency component, to the heterodyne frequency, and to all other components present in the anode current.

Consequently, these components flow through the tuned anode circuit without building up any noticeable voltage across it. The i.f. signal is passed through transformer L_3L_4 to tuned circuit L_4C_4, this circuit being connected into the control-grid circuit of the first i.f. amplifier stage.

The type of detection which takes place in the double-grid frequency conversion is neither the grid detection nor the anode detection. Let us analyse the principle of this special detection. A family of grid curves of the heptode is given in Fig. 219. These curves represent the dependence of anode current I_a upon voltage U_{g3} on the signal grid, when voltage U_{g1} on the heterodyne grid has different values (—4 v, —2 v and 0 v). The voltage on the anode and on the screen grids is maintained at a constant value. Since the frequencies of the alternating voltages at the signal grid and at the heterodyne grid are different, at certain moments these voltages coincide in phase, and at other moments they differ in phase—which is always observed during the formation of beats in any type of circuit. Let us now assume that point A is the operating point, while the voltage amplitudes of the received signal and of the signal developed by the heterodyne are, respectively, 1 v and 2 v. Then at the moment of phase coincidence the positive half-waves of voltages U_{g3} and U_{g1}, acting simultaneously, will increase the anode current up to point B, while the negative half-waves of these voltages, also acting simultaneously, will decrease the same current to point C. In the case just described, the amplitude of anode current pulsations will be much greater during a positive half-period than during a negative half-period, as shown by the anode current curve in the right-hand side of Fig. 219. After a certain lapse of time, there will come a moment when voltages U_{g3} and U_{g1} will become opposite in phase. Now the positive half-wave of voltage U_{g3} will act simultaneously with the negative half-wave of voltage U_{g1}, resulting in the decrease of the anode current to point D.

In the similar way, the positive half-wave U_{g1} will act during the negative half-wave U_{g3}, giving an increase of the anode current up to point E. The right-hand curve shows the oscillations of the anode current between the two described moments and also up to the next moment of phase coincidence. Beats of asymmetrical shape have been produced in the anode current. These beats have a component which pulsates at the intermediate frequency, shown by the thick line in the graphic drawing at the right. Thus, in the type of double-grid frequency converter described above, the detection process is performed because of the curvature of the valve characteristic and also because of the change of the characteristic itself, caused by the variation of voltage U_{g1}.

When the h.f. tuned circuits of a superheterodyne receiver are tuned by means of a ganged variable capacitor, the gang being comprised of equal-capacity sections, care should be taken that the tuned circuit of the heterodyne is always adjusted to the required higher frequency, which corresponds to the frequency of the incoming signal plus the intermediate frequency. To secure this, auxiliary capacitors C and C_5 are connected into the heterodyne

tuned circuit (Fig. 219). Capacitor C, connected in series with the main capacitor C_2, reduces the capacitance of the tuned circuit, thus providing the required higher frequency of the heterodyne. The capacitance values of C and C_5 and the inductance value of L_2 are arrived at by computation, whereupon capacitor C_5 is so adjusted that the frequency difference of the tuned circuits remains constant and equal to the intermediate frequency over the whole tuning range. Such adjustment is known as the *tracking of the tuned circuits.* In the tracking process, which is somewhat difficult, capacitors C and C_5 act as the aligning capacitors. Once the tuned circuits have

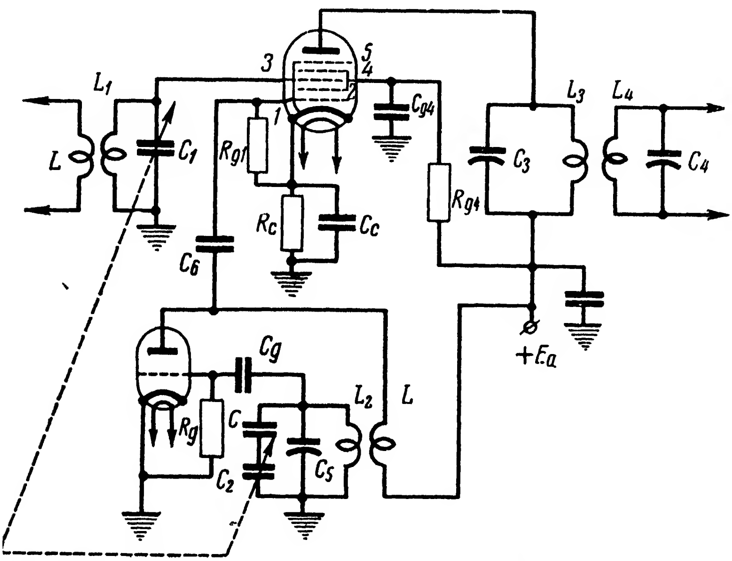

Fig. 220. The frequency converter employing a heptode mixer and a separate heterodyne valve

been adjusted to proper tracking, the aligning capacitors and the cores of the coils should not be touched, otherwise the tracing will be upset and the audibility impaired.

In a converter stage, the operating point is set on the valve characteristic by a proper adjustment of the bias, which is generally developed by an automatic biasing circuit. In some cases the stage operates at zero bias, and then the biasing circuit becomes unnecessary. The screen-grid voltage is usually fed to the valve through a dropping resistor or else from a voltage divider.

Triode-hexodes and triode-heptodes function well as frequency converters. The triode section in such valves is the heterodyne section but is separated from the mixer section, although both sections are enclosed in the same valve envelope. The heterodyne signal is applied to the control grid of the hexode or heptode. In the

case of a triode-hexode, an internal connection is provided within the envelope to conduct this signal to the grid. Since the triode and hexode (or heptode) sections generate independent electron streams, the operation of the heterodyne is practically uninfluenced by the input-signal section of such valves. Some frequency converters also employ a heptode valve, acting as the mixer, and a separate valve functioning as the heterodyne (Fig. 220). In such a converter, the heterodyne valve may employ any one of the usual oscillator circuits. Particularly good performance, in the given case, is offered by an electron-coupled pentode oscillator. The heterodyne signal is

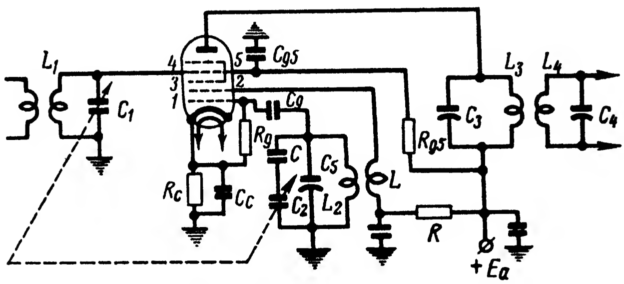

Fig. 221. An earlier type of frequency converter employing a heptode-converter valve

fed to grid *1* of the heptode valve through capacitor C_6. Resistor R_{g1} is connected in this circuit in order to drain off the electrons, which would otherwise collect on the grid of the valve.

A 6A7 valve, or a similar valve, can be used as the heptode mixer.

Fig. 221 gives the circuit diagram of an earlier type of frequency converter. In this circuit, the functions of the mixer and of the heterodyne are performed by a single heptode-converter valve type 6A8 or СБ-242. The cathode and grids *1* and *2* represent the triode (the heterodyne) section, in which grid *2* acts as the anode. Grids *3* and *5* are the screen grids, while grid *4* is the signal grid. In such a converter, the heterodyne is not completely free of the influence of the received signal; voltage changes at the signal grid affect the characteristic of the valve and its parameters, as a result of which the heterodyne frequency varies. This effect is particularly noticeable on short waves.

In order to reduce the influence of the incoming signal upon the operation of the heterodyne in a superhet receiver, a special method of frequency conversion is sometimes resorted to. In this method, one of the harmonics of the heterodyne signal is employed in the conversion process. The heterodyne is so tuned, that its 2nd or 3rd, or a still higher harmonic, has a frequency differing from the frequency of the incoming signal by the value of the intermediate frequency. For instance, if it is required to receive radio stations in the frequency range of 3,000-4,500 kc, and if the intermediate frequency is to be 120 kc, the tuning range of the fundamental frequency of the heterodyne must be from 1,040 to

1,540 kc, the third harmonic of this fundamental frequency beating with the incoming signal. The frequency range of such third harmonic will be from 3,120 to 4,620 kc, i.e., the harmonic frequency will be higher by 120 kc than the frequency of the incoming signal at any point of the tuning range of the radio receiver. This will secure the proper 120-kc i.f. signal output of the mixer at any setting of the receiver tuning dial.

A further development of this method utilises several harmonics developed by a single heterodyne. This provides for the reception of a very broad band of incoming signal frequencies, although the fundamental frequency of the heterodyne will vary over a comparatively narrow band. Again, taking the above-given example, when the fundamental frequency of the heterodyne varies within 1,040-1,540 kc, the 4th harmonic of the heterodyne (4,160-6,160 kc) will provide for the reception of signals within a 4,040-6,040 kc band. The 5th harmonic of the same heterodyne (5,200-7,700 kc) will provide for the reception of signals within 5,080-7,580 kc band. When the described method of "harmonic beating" is employed, the tuned circuits adjusted to the incoming signals have to operate over a wide band of frequencies, but the tuned circuit of the heterodyne is considerably simplified.

Since the exact tuning of a radio receiver to the frequency of the incoming signal is determined by the heterodyne frequency, the designers of radio receivers strive to secure the maximum stability of the frequency of the heterodyne stage. When a continuous wavelength coverage is not required and the receiver is called upon to operate only at several fixed waves, the heterodyne frequency is sometimes controlled by means of crystals (the same way as the master oscillator stages are controlled in modern radio transmitters). The crystal control, however, is applicable to a radio receiver only when the transmitter which is to be received by such a receiver is also crystal-controlled.

If the above described method of harmonic beating is employed in the crystal-controlled heterodyne of a radio receiver, each crystal will provide for the reception of several radio transmitting stations operating on different waves.

It should be noted that a double-grid frequency converter can employ a pentode as the mixer valve, the suppressor grid of the pentode acting as the heterodyne grid. In such a circuit, a separate valve oscillator, functioning as the heterodyne stage, will feed the heterodyne signal to the suppressor grid of the pentode.

Frequency converter valves, apart from the usual parameter indices, are also characterised by conversion transconductance S_c. The conversion transconductance is expressed in milliamperes per volt and indicates the value of i.f. current in milliamperes obtained in the anode circuit of the valve when a 1-volt signal is applied to the control grid (the signal grid) of the given valve. The value of S_c becomes greater as the voltage developed by the heterodyne increases. If this voltage is equal to zero, the value of S_c is also zero. In modern frequency-converting valves the value of S_c is considerably lower than the mutual conductance S and is usually equal to some fractions of one milliampere per volt, only in exceptional cases reaching 1-2 ma per volt. The values of S_c, given in electron valve manuals, pertain to the operation of the valve when the heterodyne voltage is normal.

Gain k of a frequency-converting stage represents the ratio of i.f. voltage, developed across the tuned anode circuit, to the signal

voltage applied to the control grid. This gain may be approximately determined by the following formula:

$$k \approx S_c R_e$$

where R_e is the equivalent resistance of the i.f. tuned anode circuit at resonance. Because S_c is always smaller than S, the amplification of the valve acting as a frequency converter is always smaller than the amplification of a h.f. amplifier or i.f. amplifier employing a similar valve.

109. INTERMEDIATE-FREQUENCY AMPLIFICATION

An intermediate-frequency amplifier (i.f. amplifier) is, in effect, a high-frequency amplifier, but it has its own peculiarities. The stages of an i.f. amplifier generally employ high-frequency pentodes. Fig. 222 gives the example of a single i.f. amplifier stage. The main peculiarity of such a stage is that it employs band filters. Usually such a filter consists of two tuned circuits which are closely and inductively coupled to each other and are tuned to the inter-

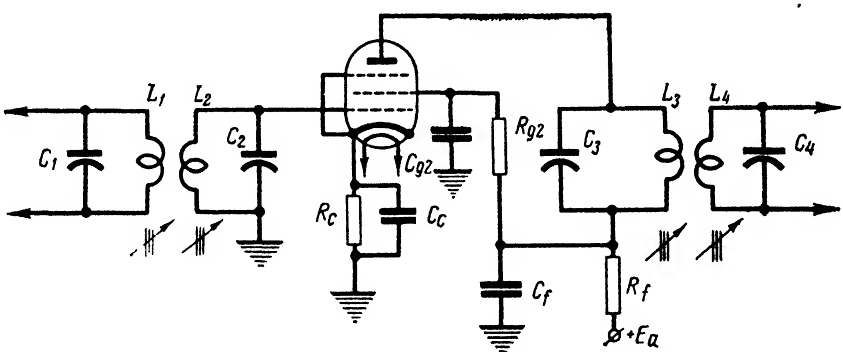

Fig. 222. I.f. amplifier stage with band filters

mediate frequency. It is this type of filter that is generally used to couple an i.f. amplifier stage to the preceding and to the following stages. Examples of such filter are given in Fig. 222 as L_1C_1-L_2C_2 and L_3C_3-L_4C_4. The tuned circuit coils of the filter represent an intermediate-frequency transformer, or i.f. transformer, as it is usually called. The coils of an i.f. transformer usually carry a variety of the honeycomb winding (for instance, the "Universal" winding).

As an alternative to i.f. transformers, band filters coupling i.f. amplifier stages sometimes employ external capacitive coupling (Fig. 223). When such type of coupling is employed, the coils are provided with closed cores made of carbonyl steel or other magneto-dielectric to obtain high value of Q of the tuned circuits. It is

impossible to obtain inductive coupling with this type of coils. Because of this, the coupling is effected by means of low-capacitance capacitor $C_{coup.}$, the value of the capacitance being selected during the adjustment of the receiver.

Trimming capacitors, simply called trimmers, are used for the exact adjustment of tuned circuits to resonance. As an alternative, the exact adjustment may be obtained by using moving-core coils. Band filters, employing the latter type of adjustment and fixed capacitors, have become very popular during the recent years.

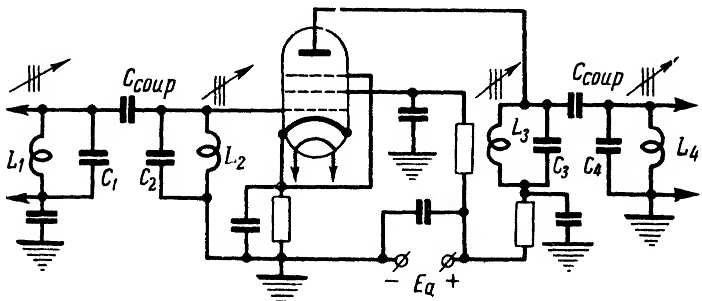

Fig. 223. The i.f. amplifier employing an external-capacitive band-filter coupling

The band filters used in i.f. amplifier stages serve to improve the selectivity of the radio receiver. In order to understand the function of such filters, let us recall the resonance characteristic of two coupled tuned circuits, which have been tuned to resonance. This characteristic, as we know, represents the dependence of the secondary tuned circuit voltage upon the frequency of the generator supplying the primary tuned circuit (Fig. 17). When the coupling is loose, the curve becomes quite sharp but possesses a wide base spreading over a great frequency range. Such characteristic is quite undesirable because the filter will not be very good at passing the sidebands of a modulated signal and, as a result, frequency distortion will occur (attenuation of the higher audio frequencies). Besides, the filter possessing such a characteristic will not be able to eliminate the interference created by other radio stations, particularly if the stations are powerful and their carrier frequencies are close to the frequency of the desired station. The filter characteristic corresponding to the optimum coupling is more advantageous, because the base of the curve spreads over a narrower frequency range (hence, the interference will be reduced), while the flat top of the curve is broader and permits the filter to pass the sidebands with greater ease. When the coupling exceeds the optimum (critical) value, the resonance curve developes side-humps, the depression between which will be the greater, the closer is the coupling.

Band filters are so designed and adjusted that the width of the resonance curve is sufficient for passing the whole frequency band of a modulated signal. In case of radio broadcasting, this frequency band must be about 9 kc (4.5 kc each side of the resonant frequency). If the receiver is designed for the reception of speech only, the filter needs to pass a frequency band of only 4-5 kc.

Fig. 203*d* gives the ideal resonance curve, the shape of which is rectangular. Such a shape corresponds to the condition when the filter passes with equal ease all the sideband frequencies, sharply cutting off the interference caused by other radio stations operating on frequencies which lie just outside the vertical sides of the curve. As we already know, the ideal resonance curve is but an assumption and cannot be attained by any practical means. The best we can do is to use as many band filters as possible in a radio receiver, and then the shape of the resulting practical resonance curve will approach closer and closer the ideal resonance curve. When a depression appears between the humps of the resonance curve, such a depression is sometimes eliminated by connecting a single tuned circuit instead of the band filter in one of the i.f. stages. Since the characteristic of a single tuned circuit usually has a "single-hump" shape, a compensation will take place and the shape of the resulting resonance curve of the amplifier will look like the curve shown in Fig. 224.

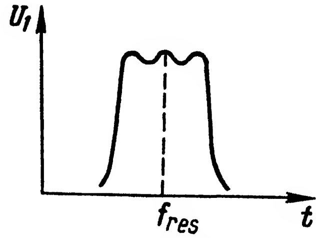

Fig. 224. A "triple-hump" resonance curve

Band filters are always enclosed in shields and are tuned to resonance by turning the adjustment screws of trimmers or coil cores. Once adjusted to resonance, these screws should not be touched, otherwise the tuning will be thrown off, resulting in a sharp decrease of amplification and in the distortion of the correct shape of the resonance curve. The repair and adjustment of band filters may be conducted only by an experienced radio technician with the help of appropriate measuring devices.

The intermediate frequency value is usually so selected that it falls into that part of the radio spectrum in which radio transmitters only very seldom operate. Should the intermediate frequency be set to the same value as the carrier frequency of some radio transmitting station, the radio receiver placed close to such station will continuously reproduce the signals of this station — regardless of the setting of the receiver tuning dial. In this case, the signals of the offending station will be directly reaching the i.f. amplifier through various parasitic capacities which usually exist between the receiving aerial and various parts of the receiver. When the intermediate frequency of a receiver has to be set close to the frequency of some operating radio station, the type of interference just described

may be eliminated by connecting a special filter into the input circuit of the receiver. The filter is often connected right into the aerial and, when tuned to the intermediate frequency of the receiver, will keep the offending signal away from the radio set.

It is not advisable to select too low a value of the intermediate frequency f_{if}. If the intermediate frequency is equal only to some dozens of kilocycles or so, the reception will be greatly interfered with by radio stations whose carrier frequencies differ from the carrier frequency of the required radio station by $2f_{if}$ (the case of image interference). Too high a value of the intermediate frequency is also undesirable because, in this case, i.f. amplifier stages will provide smaller gain and poorer selectivity.

I.f. amplifier stages, as a rule, employ automatic biasing arrangements. The screen-grid voltage is fed through a dropping resistor or else from a divider. Decoupling filters are frequently used in i.f. amplifiers in order to eliminate parasitic coupling through the common anode circuits. The gain of an i.f. amplifier stage, employing a good high-frequency pentode, can be as high as 50 and even 100.

Some superheterodyne receivers use special band filters in their i.f. amplifiers, such filters consisting of more than two tuned circuits to improve the selectivity of the set. On the other hand, ultra-short-wave receivers use single tuned circuits instead of the usual i.f. transformers.

The selectivity of the i.f. amplifier may be greatly increased through the use of an extremely high Q piezo-electric series-resonant circuit. The piezo-electric quartz crystal, together with its coupling arrangement, is generally known as a crystal filter. Apart from its very high value of Q, the crystal filter also possesses an exceptionally high stability. A filter of this type has an extremely sharp resonance curve, i.e., it is capable of passing such a narrow frequency band, the width of which is equal to only some dozens of cps. This sharply cuts down the interference set up by other radio stations in a radio receiver.

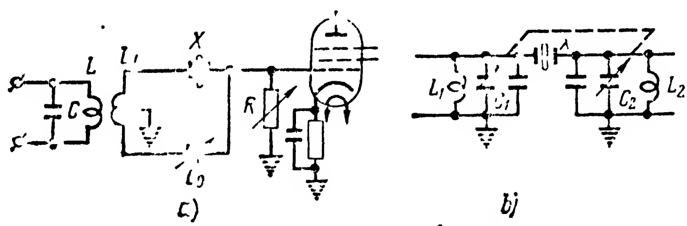

Fig. 225. Crystal filter circuit diagrams

The simplest circuit diagram of a crystal filter is shown in Fig. 225*a*. Tuned circuit *LC* is adjusted to the intermediate frequency and is connected into the anode circuit of the frequency converter. The centre-tap of coil L_1 is earthed. The voltages set up in the two halves of the coil are equal in value but opposite in phase. These voltages are fed to the grid of the valve through crystal *X*, which is designed for the intermediate frequency, and through capacitor C_0, whose capacitance is equal to the capacity of the crystal holder. If the

signal frequency does not fall within the limits of the frequencies which can be passed by the crystal, the latter functions only as a capacitor, whose capacitance is equal to C_0. Under such operating condition, the two alternating voltages fed to the grid of the valve are equal to each other and are mutually cancelled. However, when the signal frequency falls within the limit of the frequencies that can be passed by the crystal, the latter begins to resonate and functions as a series-resonant circuit. In this case, the crystal impedance sharply drops, and the voltage applied to the valve grid through the crystal will be higher than the voltage fed through capacitor C_0. These voltages now no longer compensate each other and the process of reception begins.

Resistor R functions as a load resistor and exerts a strong influence upon the bandwidth passed by the crystal filter. The lower the value of R, the greater is the bandwidth. If this resistor is made variable (as is always the case), it can be so adjusted that the bandwidth can be varied from some dozens to several hundred cycles per second. The narrower bandwidth is used on radio telegraphy and gives an exceptionally high selectivity. If the bandwidth is expanded to about one thousand cps, the reception of radio-telephone signals becomes possible. True, the telephone signals will be distorted to a considerable degree, but, then, the interference, caused by stations operating on nearby frequencies, will be greatly reduced, in comparison with the case when no crystal is used in the receiver. In some cases, resistor R is replaced by a tuned circuit, adjusted to the intermediate frequency, and possessing (besides the usual capacitance and inductance) also some changeable ohmic resistance. In such a circuit, the adjustment of the bandwidth is obtained by varying the said ohmic resistance of the tuned circuit. The bandwidth can also be varied within certain limits by changing the capacitance of capacitor C_0.

Fig. 225*b* gives another example of a crystal filter, where the crystal is used as the capacitor in the system of external capacitive coupling. In this circuit, the bandwidth that can be passed by the crystal is adjusted by detuning the oscillatory circuits both sides of the resonance frequency (the frequency of the crystal). Such detuning is performed by a special gang of variable capacitors C_1 an C_2. This gang is so designed that, when the common rotor shaft of the two capacitors is turned, the capacitance of one of these capacitors increases, while the capacitance of the other decreases.

Other, more complex, circuit arrangements of crystal filters are also used, and in some of these circuits two and even more crystals are employed.

110. THE SECOND DETECTOR, BEAT-FREQUENCY OSCILLATOR AND LOW-FREQUENCY AMPLIFICATION

Superheterodyne receivers generally employ diode detectors. Anode and grid detectors are only occasionally used in receivers of this type. This is attributed to the fact that a superhet provides high gain and its second detector is supplied with comparatively trong signals. Only the diode detector can provide practically distortionless detection on strong signals, and this is the reason why it is almost universally used in all types of superheterodyne receivers. A typical circuit diagram of a diode detector is given in Fig. 226*a*.

In such a circuit, the signal developed in the output transformer of the last i.f. amplifier stage is fed to the detector. R_1 and R_2 are the load resistors of the detector, these two resistors being connected in series and constituting one common load resistor. The audio frequency is taken off from across resistor R_2 and is delivered to the input

circuit of the first low-frequency amplifier through a potentiometer, generally known as gain control (R). The described detector circuit is very often employed not only because it provides good detection, but also because it keeps high-frequency oscillations away from the low-frequency amplifier, and also because it reduces distortion.

Modern superheterodyne receivers make a wide use of special complex valves, such as double diode-triodes and double diode-pentodes. It is a general practice to employ one of the diode sections

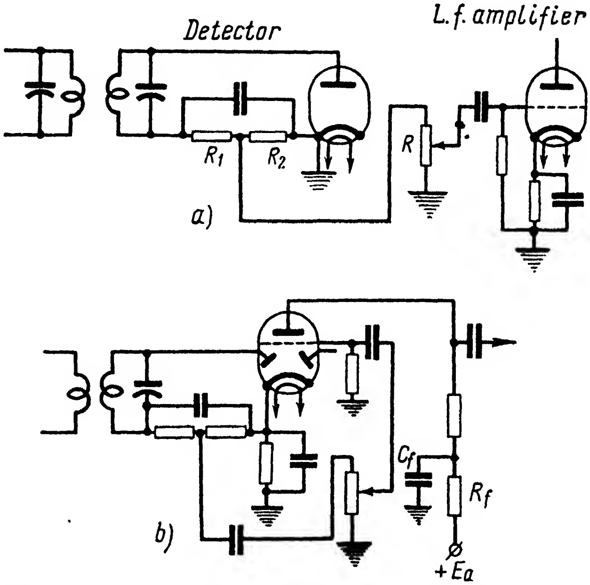

Fig. 226. The circuit diagram of a detector and of its connection with l.f. amplifier

of such a valve for the purpose of detection, the triode or pentode section of the valve serving as the first low-frequency amplifier (Fig. 226*b*). Pentode section in these valves may also be used as an i.f. amplifier, the diode section being employed for detection, but this circuit arrangement is rather rare.

As noted earlier, the electron valves used in radio receivers develop internal noise interfering with the radio reception. In order to reduce the level of such noise, a small fixed negative voltage bias is sometimes applied to the anode of the diode detector valve. This voltage cuts off the diode current set up by the noise voltage and, therefore, prevents the noise from reaching the output circuit of the radio receiver. However, when the signal of a radio station reaches the negatively-biased diode detector, such a signal, having an amplitude much higher than that of the noise voltage, causes current in

the diode. In such a case, the intelligence contained in the signal is reproduced by the receiver practically free of any accompanying noise. The negative voltage (the bias) may be applied to the anode of the diode detector from the self-biasing resistor connected into the anode circuit of some other valve of the receiver. The detector itself cannot be self-biased, because, when no signal is present, the diode current is zero.

Beat-frequency oscillators are employed by superheterodyne receivers only for the reception of c.w. telegraph signals. (A beat-frequency oscillator is also useful when searching for weak carriers of radio-telephone stations. —*Translator's note.)*

Most types of beat-frequency oscillators employ inductive feedback and triode or pentode valves. The oscillator may be inductively or capacitively coupled to the tuned circuit of the detector. A switch provided in the radio receiver stops the beat-frequency oscillator when a modulated signal is tuned in. The tuned circuit of the beat-frequency oscillator is adjusted by means of a trimmer or a moving coil core, the oscillator being made to operate on a constant frequency, differing from the i.f. frequency by about 1,000 cps. An example of a beat-frequency oscillator circuit coupled to the 2nd detector in a radio receiver is shown in Fig. 227. Straight-amplification receivers and superheterodynes employ the usual low-frequency amplifiers, whose operating principles and peculiarities were already discussed under Chapter VII.

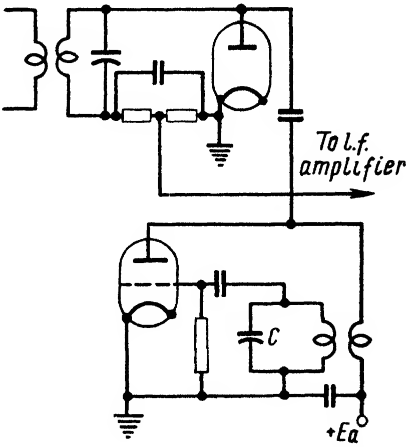

Fig. 227. A beat-frequency oscillator, coupled to a diode detector

A detector stage using a triode, or a pentode, is, in effect, also the first low-frequency amplifier valve. Because of this, when a radio receiver is also designed to reproduce gramophone records, the pickup is connected to the control-grid circuit of the detector valve. In order to reduce distortion during the gramophone operation, a negative bias is applied to the control grid. Fig. 228*a* shows a simple circuit, in which the automatic bias voltage is fed to the valve grid only when the radio set operates from the pickup. If the pickup is of the piezo-electric type, it should be shunted by a resistor in the given circuit. If the receiver employs a diode detector, the pickup is connected in accordance with the circuit diagram given in Fig. 228*b*.

The first low-frequency stages in a radio receiver usually use resistance-coupled triodes or pentodes. The transformer or choke

coupling is seldom encountered in such stages. The output low-frequency stage employs a higher-power valve, which may be a triode, a pentode or a beam tetrode. When the output stage is designed to drive an electrodynamic loudspeaker, such a stage is always transformer-coupled to its load. Only the simplest receivers employ direct coupling, to drive magnetic speakers or else earphones. Radio receivers are often provided with separate output terminals for the connection of an auxiliary external loudspeaker.

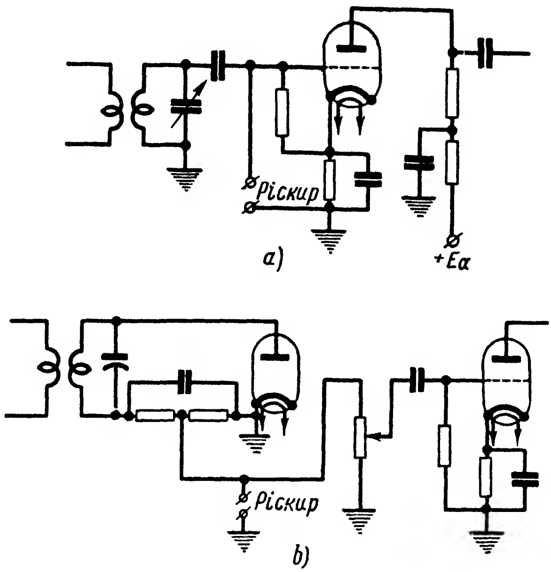

Fig. 228. The connection of a pickup to a radio receiver

Receivers of a more elaborate design employ push-pull output stages. The low-frequency stages of radio receivers incorporate all the features of low-frequency amplifiers, discussed under Chapter VII. Such features include automatic biasing circuits, various methods of feeding screen-grid circuits, decoupling circuits, negative feedback, phase-inverting circuits, etc.

111. REGENERATIVE DETECTORS

Although the superheterodyne receiver, because of its outstanding merits, holds the leading place in the field of modern radio receivers, there also exist other types of receivers which deserve consideration. We have already studied the straight-amplification receiver and

know its peculiarities. Now we shall discuss a special version of such receivers, where the effect of the so-called regeneration is used.

As a matter of fact, the detector stage of most modern straight-amplification sets employs regeneration.

In the regenerative detector a high-frequency feedback circuit is provided between the anode and grid circuits, such a circuit resembling the usual feedback arrangement of a self-excited valve oscillator. The circuit of the regenerative detector with an inductive type of coupling is shown in Fig. 229*a*. In the regenerative radio receiver the feedback is always made variable and capable of smooth adjustment. The principle of regeneration may be explained as follows.

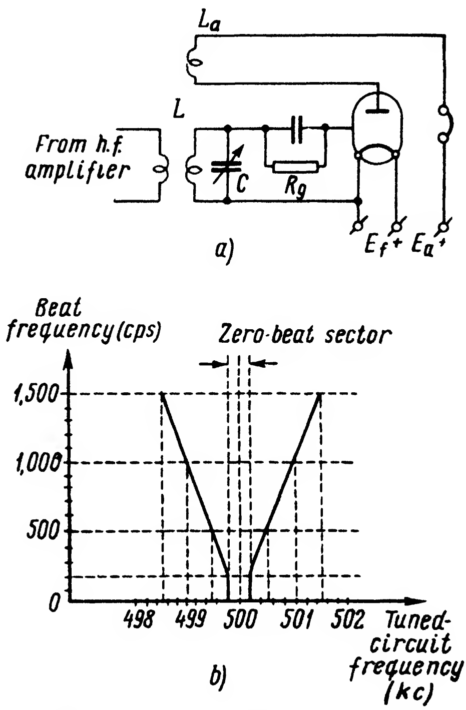

Fig. 229. The circuit diagram of a regenerative detector with inductive feedback, and a graphic representation of beat-frequency tone changes during the tuning of such detector

The anode current of a valve detector is a pulsating current and consists of three components — the d.c. component, the low-frequency component and the high-frequency component. The shape of the high-frequency component corresponds to the shape of the modulated signal being received. This component flows through feedback coil L_a and induces an alternating voltage in the tuned-circuit coil L. If the ends of coils L and L_a are connected correctly, then the voltage induced in coil L will coincide in phase with the signal voltage and will be added to it. As a result, the alternating voltage applied to the grid of the valve will be increased. But then the amplitude of the high-frequency component of the anode current will also increase. This, in turn, will increase the voltage induced by the high-frequency component in the tuned-circuit coil. The grid voltage will be increased even more. This, in turn, will again produce an increase of the high-frequency component of the anode current. Owing to the presence of feedback, the alternating grid voltage will again be raised and the process of building up the amplitude of the oscillations will continue. Of course, it will continue only up to a

certain limit. As the amplitude of the oscillations increases, the energy losses in the ohmic resistance of the tuned circuit also increase. As we already know, the building up of the amplitude of oscillations can continue only while the energy added to the tuned circuit by means of the feedback is greater than the energy lost in the circuit. As soon as the losses of energy become equal to the added energy, the building up process will stop.

This process, pertaining to the building up of the amplitude of high-frequency oscillations, becomes possible only when the value of feedback is sufficient in quantity and correct in phase, i.e., only when coils L and L_a are sufficiently close to each other and are connected correctly. If the coil connections are wrong, the voltage induced in the tuned circuit by coil L_a will be opposite in phase to the signal voltage and will not amplify the oscillations in the tuned circuit but will have the opposite effect of attenuation on these oscillations.

Assuming that the coils have been connected correctly, we note that when these two coils L_a and L are moved up closer to each other, i.e., when the value of feedback is increased, the amplification will also be increased, but up to a certain point only. At this point the feedback becomes so large that the stage begins to produce its own oscillations and, thus, becomes, in fact, an oscillator stage.

The value of feedback at which the stage begins to oscillate is known as the threshold of oscillation.

When a radio receiver generates its own oscillations, the latter are added to the oscillations of the incoming signal and a bad distortion of the received intelligence takes place. If the incoming signal is unmodulated (a c.w. signal), beats are produced in the receiver as a result of addition of the two types of oscillations. The frequency of such beats depends upon the frequency difference between the incoming signal and the local high-frequency signal generated by the receiver.

The frequency of the incoming signal is constant, while the frequency of the local signal is determined by the parameters of the tuned circuit of the oscillating stage of the radio receiver. If one of such parameters is varied (for instance, if the capacitance of the tuned circuit is varied), the frequency of the local signal will change. As a result, the beat-frequency, produced by the mixing of this signal with the incoming signal, will also change. Consequently, if the tuned circuit of the radio receiver employs a variable capacitor, a slight rotation of the rotor of such capacitor will give beats of various frequencies. After these beats have been detected, an alternating component of the beat-frequency will be produced in the anode circuit of the output of the stage. If the beat-frequency is within the limits of the audible range, the alternating component will produce a sound when flowing through a pair of earphones or a loudspeaker.

When the tuned circuit of the receiver is adjusted to exact resonance with the frequency of the incoming signal, no beats will be produced, because the difference between the two frequencies is equal to zero. This is known as the zero-beat condition, and the non-existent beats, in this case, are called zero beats.

Zero beats are an indication that the tuned circuit is adjusted to exact resonance with the frequency of the external signal. It is interesting to note that the zero-beat condition can be also attained at a slight detuning of the circuit in respect to the incoming signal. The zero-beat condition, in this case, is attributed to the following: when the receiving tuned circuit is only slightly detuned from the frequency of the incoming signal, this signal acts as an external excitation and makes the tuned circuit oscillate not at its natural frequency but at the frequency of the signal.*

This frequency-locking effect creates a certain sector in which the zero-beat condition prevails. The stronger the incoming signal, the broader will be such zero-beat sector. Beyond the limits of this sector, the receiver stage begins to oscillate at its natural frequency. This frequency is different from the frequency of the incoming signal, and, as a result, the two frequencies beat with each other, producing audible beats which are reproduced by an earphone as a continuous sound of sustained pitch. The larger the detuning of the receiver circuit from the incoming signal, the higher will be the beat frequency and the higher will become the pitch of the resulting sound. When the detuning is sufficiently large, the resulting signal will pass the limits of audibility and, entering the supersonic frequency range, will no longer be perceived by the ear.

Having grasped the above explanation, the reader will easily understand the following.

When the tuning of the regenerative receiver, operating in the oscillatory condition (because of the action of the feedback circuit), is so adjusted that the set is brought closer and closer to the condition of resonance with an incoming carrier signal, first a high-pitched whistle will be heard. On further rotation of the tuning control, the pitch of the sound will be gradually lowered, the sound stopping altogether at a certain low-audio frequency, just before the point of exact resonance is reached (the zero-beat sector). As the tuning control is advanced further, and the point of exact resonance is passed, the above-described effects are repeated, but in the reverse order; first, a low-frequency sound will appear, then the sound will become more and more high-pitched, and, finally, it will no longer be heard, as the tuning control continues to detune the receiver oscillatory circuit further and further away from resonance with incoming carrier signal.

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* The effect just described is known as forced synchronisation or frequency locking.

The described process is shown graphically in Fig. 229*b*, where the beat frequencies are marked on the Y-axis, while the frequencies of the oscillating receiver are marked along the X-axis. The given graphic representation takes a case when the frequency of the incoming signal is equal to 500 kc. In this case, audio-frequency beats appear only when the receiver circuit is detuned by at least 200 cps from the incoming signal. Thus, in the given example, the width of the zero-beat sector is equal to 400 cps.

When it is required to receive a radio telephone signal on the regenerative set, the latter must not oscillate, otherwise the whistle produced by the beats will badly mar speech and music. The reception of the modulated signal could be effected by adjusting the oscillating detector to the zero-beat sector, but this sector is noted for its notoriously poor stability. Because of this, when a modulated signal is to be produced by a regenerative receiver, the value of feedback in such a receiver is reduced until the set no longer oscillates, although it remains close to the threshold of oscillation.

On the other hand, the detector of the regenerative receiver must be in the state of oscillation if it is to pick up a c.w. telegraph signal. If the detector stage is not adjusted to the point of oscillation in the given case, the c.w. signal will produce only unintelligible clicks in the earphones. And only when the oscillatory-detector-condition is set, will the earphones reproduce dots and dashes of the telegraph code as short and long audible tones. Such tones will result from the beating of the incoming carrier signals with the signal of the oscillating detector, the most pleasant pitch of the tones being obtainable by the adjustment of the tuned circuit of the receiver heterodyne. Radio operators usually adjust their sets to produce a 1,000-cps tone, because the ear is most sensitive to this frequency. As we already know, such 1,000-cps tone is obtained when the tuned circuit of the receiver heterodyne is detuned by 1,000 cps from the frequency of the incoming signal. The circuit may be detuned by this value either side of the resonance.

As follows from the above discussion, the loudest reception of radio-telephone signals is obtained near the threshold of oscillation, when the detector is not yet oscillating but is close to the oscillatory condition. At the same time, the best reception of c.w. signals is secured only when the detector is adjusted so that the threshold of oscillation is just passed, i.e., when the feedback value is just sufficient to commence the oscillations. It should be noted that the feedback value should not be allowed to increase beyond this point, because the loudness of the telegraph signals will be reduced if the amplitude of the locally-generated signal is high. The value of feedback in a regenerative receiver is controlled by means of a separate control, generally known as the regeneration control. Fig. 230 shows the various positions of the regeneration-control handle, securing

respectively the most advantageous values of feedback for the reception of telephone and telegraph signals.

The circuit shown in Fig. 229 is seldom used, because shifting of the feedback coil in the proximity of the tuned-circuit coil exerts a strong influence upon the inductance and capacitance of the tuned circuit and changes its frequency. This makes the detector stage difficult to tune. As a matter of fact, the tuning scale of such a detector cannot be accurately calibrated. The regenerative detector circuit employing a variable capacitor to control the feedback is more rational and is, therefore, more frequently used. A circuit of this type is shown in Fig. 231*a*. In such a circuit, the d.c. component and the audio component of the anode current pass through high-frequency anode choke *Ch*, this choke blocking the high-frequency component and not letting it pass from the valve anode to the low-frequency output circuit. Thus, the high-frequency component is forced to flow only through the feedback coil, known as the tickler coil, and through variable capacitor C_a, the latter acting as the regeneration control. Coils L_a and L are placed close to each other and are fixed in this position. When the regeneration control

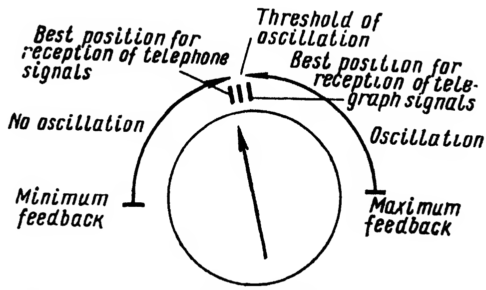


Fig. 230. Various positions of the regeneration control handle

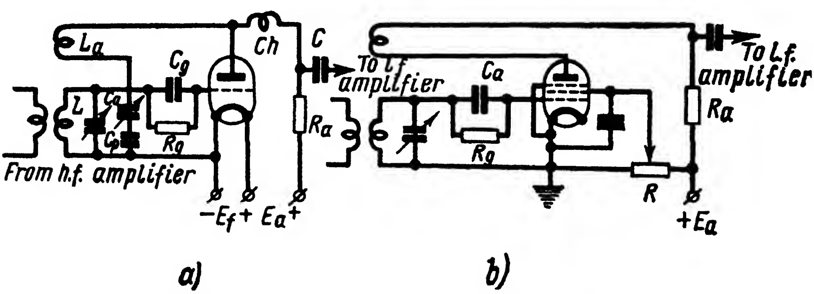


Fig. 231. Regenerative stage circuits employing a capacitor-type regeneration control *(a)* and a potentiometer-type regeneration control *(b)*

capacitor C_a is rotated in the direction effecting an increase of its capacitance, the high-frequency current, flowing through coil L_a, increases and the feedback is intensified. The capacitance of C_a may be anywhere between 200 and 500 pf and this capacitor can employ a solid dielectric. Sometimes a fixed 1,000-2,000 pf capacitor C_p is connected in series with capacitor C_a, serving as a protective

capacitor to prevent the short-circuit of the anode power supply in the event of an accidental contact between the rotor and stator plates of capacitor C_a. Coil L_a usually consists of a smaller number of turns than the tuned circuit coil and may be wound with any grade of thin wire. If the two coils L_a and L are placed on the same coil form, the parasitic capacitance can be reduced by connecting to the anode and to the grid those ends of the coils which are most remote from each other. In order to avoid too high a value of feedback, it is recommended to connect a small capacitor (20-50 pf) between the anode and cathode of the valve. Then the high-frequency current will be partly branched off through this capacitor and will not flow exclusively through coil L_a. High-frequency choke *Ch* consists of several hundreds of turns and its distributed capacitance must be made as low as possible. The choke winding is sometimes sectionalised for this purpose.

Regenerative detectors frequently use pentode valves providing a high degree of amplification. In this case, the regeneration control can employ a variable capacitor, as shown in Fig. 231*a*, or, else, a potentiometer varying the screen-grid voltage (Fig. 231*b*).

Proper operation of a regenerative receiver is obtained only by a *smooth* and gradual approach to the regeneration stage. Such smooth approach is a necessity because the amplification reaches its maximum value just near the threshold of oscillation. If the circuit parameters ensure a smooth passage of the characteristic of regeneration, a decrease of feedback, after the detector had begun to oscillate, stops the oscillation at the same value of feedback at which the oscillation was commenced. Only one threshold of oscillation is obtained, such threshold being definite and stable. Weak signals may be received, and such signals will be stable in the direct proximity of the threshold of oscillation, where the amplification is at its maximum value.

An opposite effect is observed in some regenerative sets, where the approach to the regeneration point is *sharp* and where the *pulling-in* of the regeneration takes place. These effects are evidenced as follows. The oscillation begins with a click in the earphones, cannot be stopped when the regeneration control is slightly retarded, and ceases again with a click after the control has been set to a much lower value of feedback than that at which the oscillation had commenced. In the case just described, two unstable oscillation thresholds are observed and the approach to a point just short of regeneration becomes impossible. The sharp regeneration can be eliminated by reducing the anode voltage of the detector valve and by selecting the grid resistor of correct value. The smooth approach to the point of regeneration depends upon the position of the operating point on the grid characteristic of the anode current. It is important that this point is located on the steepest part of the curve, i.e., on the linear part. Since the anode detection calls for the

location of the operating point at the lower bend of the curve, anode detectors should not employ regeneration.

The best detector action and smooth approach to the point of regeneration in battery-powered radio receivers are secured not only by the selection of the correct value of resistor R_g, but also by determining whether this resistor should be best connected to the positive or to the negative terminal of the filament battery. If it is connected to the negative terminal, the approach will be smoother. On the other hand, connection of this resistor to the positive terminal will increase the loudness of signals. The best compromise is sometimes found by connecting R_g to the slider of a potentiometer shunting the two terminals of the filament battery, and finding the working point by trial and error.

As we already know, the grid detector is not only a detector but also a low-frequency amplifier. When regeneration is incorporated into such a detector, the stage will provide a high value of h.f. gain. Hence, the regenerative detector is very sensitive, particularly to weak signals. The weaker the signal being received, the higher will be the amplification of the detector stage. On such weak signals the gain of a regenerative receiver can be as high as several thousands. On the other hand, when such a receiver is tuned to the carrier wave of a nearby high-power radio transmitting station (which can be heard sufficiently well without any regeneration), the feedback circuit will provide but a low gain, in most cases insufficient for loudspeaker operation.

The incorporation of regeneration in a straight-amplification receiver improves the selectivity of the set. The selectivity reaches its maximum value at the threshold of oscillation, where the frequency band passed by the receiver becomes narrow. When the regenerative receiver is tuned to exact resonance with the frequency of the radio station, the reproduction will sound muted because of the cutting-off of the sidebands. The high selectivity of the receiver is impaired by high-power stations which can be heard over a considerable part of the tuning dial badly interfering with the reception of other stations. This shortcoming of the regenerative set can be overcome to a certain degree by incorporating h.f. amplifiers in the receiver.

H.f. amplifiers are desirable in all regenerative receivers anyway, because of the following unpleasant property of the regenerative detectors. When such detectors are made to oscillate on the reception of c.w. signals, they are converted into low-power transmitters. Under such conditions, the aerial of the regenerative receiver radiates radio waves and produces interference (whistles, howls, etc.) in other receivers in the given locality. Such effect can be counteracted by including at least one h.f. amplifier stage between the aerial and the regenerative detector. This done, the detector is isolated from the aerial and the latter will not transmit the locally-generated oscillations.

There was a time when regenerative receivers were quite popular. At present, however, only the cheapest sets employ regeneration. It is possible to design a straight-amplification receiver with high sensitivity and selectivity only by providing the set with many h.f. amplifier stages. A receiver of such type, however, would be prone to develop parasitic oscillations. Moreover, the cost of the set would be prohibitive, while its construction and adjustment would be quite complicated. The cost of any radio receiver is in a great measure determined by the number of high-frequency amplifier stages the set employs. Apart from the question of cost, engineering difficulties arise if the number of such stages is excessive. Two or even three tuned circuits can be ganged to operate from a common handle. But the proper tracking required for the ganging becomes almost impossible with a greater number of h.f. stages. Besides all this, h.f. amplifier stages provide only low gain on short and medium waves, because the impedance of tuned anode circuits becomes low on such waves. The straight-amplification receiver is not easy to tune; during the tuning procedure it becomes necessary to rotate simultaneously the tuning knob and the knob of the regeneration control. Weak signals may be received only when the regeneration control is set at the very threshold of oscillation. This point is difficult to find and the radio operator becomes proficient at handling the regenerative receiver only after a great deal of practice. All the listed defects pertain only to straight-amplification regenerative sets, and none of them — to the superheterodyne receiver, which is the most popular type of radio receiver today.

Some superheterodyne sets, particularly those of simpler construction, employ a regenerative grid detector instead of the beat-frequency oscillator. Because of regeneration, such receivers can reproduce not only radio-telephone signals, but also c.w. telegraph transmissions. The feedback circuit, when brought close to the point of oscillation, also provides a certain amount of gain when a set of this type is tuned to a modulated carrier wave.

112. SUPERREGENERATIVE RECEIVERS

Superregenerative receivers employ a feedback circuit, like the regenerative receivers, but operate under the condition of periodically-interrupted oscillation. These receivers, commonly known as superregenerators, are chiefly used on ultra-short waves for the reception of radio-telephone signals, as well as for the reception of modulated c.w. messages. The operating principle and the peculiarities of the superregenerator are quite different from those of the regenerative receiver, which we have already discussed.

The regenerative receiver is most sensitive to radio-telephone signals when its feedback is set to the threshold of oscillation. Under such operating condition, the receiver is extremely sensitive to the incoming signals, particularly to weak signals, but the reception is unstable. The slightest change of supply voltages or the slightest variation of receiver tuning or of the frequency of the incoming

signal — and the regenerative receiver either begins to oscillate (which badly distorts the signal), or else its sensitivity is sharply reduced.

The same regenerative receiver provides a better stability of reception on radio-telegraph signals, because in this case the set is kept under the condition of slight oscillation. The tone of the telegraph signal being received is determined by the beat frequency, this frequency being equal to the difference between the frequency of the signal and the frequency of the locally-generated oscillations. In this case, slight changes of the operating condition of the set do not, as a rule, stop the locally-produced oscillations and the set continues to operate normally. True, in the latter case, the tone of the signal will sometimes vary, because all changes of the locally-generated oscillation frequency will act to change the beat frequency, i.e., the resultant frequency. The volume of the signal being reproduced will also change in some cases. All this is tolerable up to a certain extent. However, the oscillating condition of the regenerative receiver is not permissible on reception of modulated signals, as the beats will be superimposed upon the radio-telephone intelligence and will cause bad distortion.

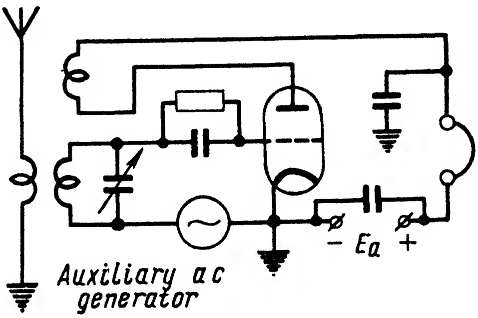


Fig. 232. The basic circuit diagram of a superregenerative radio receiver

The described shortcomings of the regenerative receiver are not encountered in the superregenerator. The latter type of receiver is maintained in the state of oscillation on the reception of radiotelephony, but the signal comes through clear because the locally-generated oscillations, in this case, are periodically interrupted at a frequency which is out of the range of audibility. Because of this, the superregenerator provides a much more stable reception of modulated signals than does the regenerative receiver. Moreover, it also provides an extremely high sensitivity. The gain that can be obtained from a single superregenerative stage reaches several hundred thousands on the reception of weak signals. The selectivity of the superregenerator is rather low. This allows us to employ this receiver for the reception of signals radiated by self-excited oscillator-transmitters which incorporate no frequency-stabilising facilities. The shortcoming of the superregenerator is the loud noise (hiss) which it develops when no signal is being received. The incoming radio signal, if it is not too weak, will stop this hiss.

The simplified explanation of the action of the superregenerative receiver is as follows. Assume that we have a regenerative receiver (Fig. 232), in which the value of feedback is so set that, at a small negative bias at the grid of the valve, the valve generates the local oscillations, while, when the bias is made more negative, these oscillations stop. Now assume, that we incorporate in the receiver circuit a component which is not used in the circuits of the usual regenerative sets — an auxiliary a.c. generator, shown in the drawing. The generator develops an alternating voltage of a certain frequency, which is considerably lower than the frequency of the local oscillations generated by the tuned circuit of the receiver. Assume that this alternating voltage is applied between the grid and cathode of the valve. When a positive half-wave of this auxiliary voltage is applied to the grid, the operating point on the grid characteristic of the valve is shifted to the position of the higher mutual conductance and the stage begins to generate the local high-frequency oscillations. During the next half-wave, the a.c. generator polarity reverses, the grid becomes more negative in relation to the cathode, the operating point is shifted to the section of the curve with a smaller slope, and the locally generated oscillations are stopped. Thus,

the generation of the local h.f. oscillations in the superregenerative receiver is interrupted at a lower auxiliary frequency. This frequency is called the quenching frequency because it serves to interrupt (to quench) the local h.f. oscillations.

When no incoming radio signal is being received by the superregenerator, the set produces the local h.f. oscillations (during the positive half-waves of the quenching voltage) under the influence of electric fluctuations. The latter name stands for very weak electric pulses which exist in any type of electric circuit because of the haphazard thermal movement of the electrons in every conductor.

Electric processes taking place in the superregenerative receiver in the absence of incoming signal are illustrated in Fig. 233. The quenching-frequency voltage, the shape of which is assumed to be rectangular for the sake of simplification of the explanation, is shown in Fig. 233*a*. Because of the rectangular

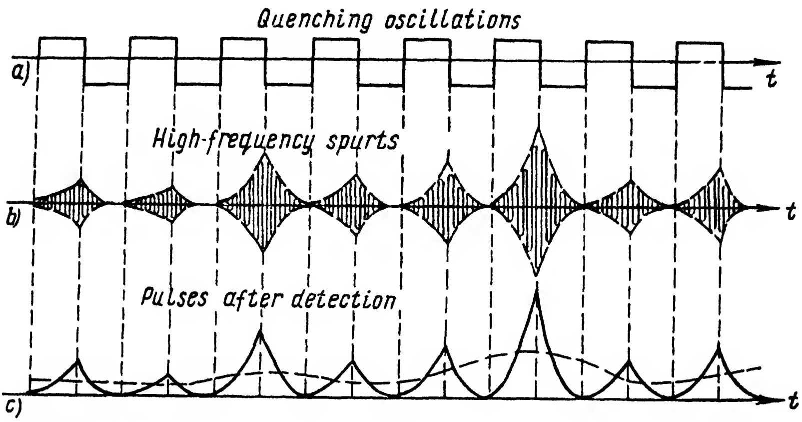


Fig. 233. A graphic representation of processes taking place in a superregenerative detector in the absence of external signals

shape of this voltage, the negative bias at the grid of the valve is low and remains constant during the positive half-waves. Because of this, high-frequency oscillations take place during such half-waves and the amplitude of these oscillations is built up. As the polarity of the quenching voltage at the grid of the valve swings in the negative direction, the grid immediately acquires a high negative charge, the conditions necessary for the self-excitation no longer exist, and the oscillations die out.

If the shape of the quenching voltage is not rectangular but sinusoidal, the operating principle of the receiver will remain the same, but the process of superregeneration will become more complex, as the gradualness of voltage change at the grid of the valve will cast a different influence on the processes of growth and decay of oscillations.

High-frequency spurts appearing in the superregenerative receiver are shown in Fig. 233*b*. These spurts appear and grow in amplitude during each positive half-wave of the quenching voltage, damping out and disappearing during the negative half-waves. The stronger the initial pulse, caused by the electric fluctuations, the greater will be the amplitude of the generated oscillations. Since the electric fluctuation pulses have random magnitudes, the high-frequency spurts also have different amplitude values; and thus, in effect, they carry a haphazard modulation. On detection of such oscillations, pulses of varying magnitude are obtained, these pulses following one another at the frequency of quenching (Fig. 233*c*). Such pulses cannot be heard in the earphones, because

the quenching frequency is a supersonic frequency. The average current value of these pulses, shown by the broken line, also changes in a haphazard way. However, this change is slower and produces a hissing noise in the earphones connected at the output of the superregenerative receiver.

When the receiver is tuned to a radio signal whose level is lower than the level of the electric fluctuation pulses, the described process will not be changed to any considerable degree. The earphones will continue to reproduce the hiss, the latter blanketing the incoming signal. However, when the level of the signal is higher than the level of the electric fluctuation pulses, the process will be performed differently (Fig. 234). A graphic representation of the quenching frequency is given in Fig. 234*a*. Fig. 234*b* shows the modulated oscillation of the incoming signal. The high-frequency spurts now arise, not because of the weak fluctuation pulses, but due to the action of the stronger incoming signal. The

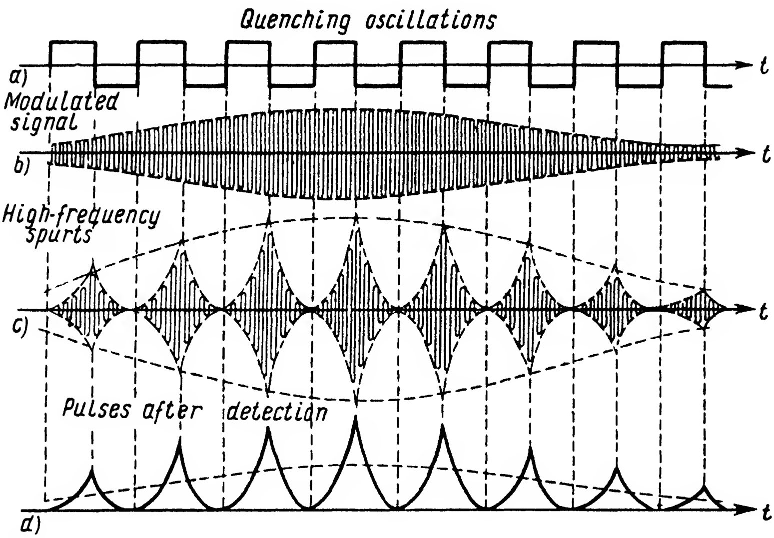


Fig. 234. A graphic representation of processes taking place in a superregenerative receiver during the reception of modulated signals

greatest amplitude of oscillations in these spurts is determined by the amplitude of the signals, i.e., the amplitude variations of the spurts follow the amplitude variations or, in other words, the modulation, of the incoming carrier wave (Fig. 234*c*). The result of detecting such signals is given in Fig. 234*d*. The average value of the received pulses varies at the frequency of the modulation, and, because of this, the earphones will now reproduce the sound being transmitted. Since the high-frequency spurts in this case are not caused by the fluctuation, the hiss will be suppressed and will not be audible even when the carrier is not modulated. Thus, the incoming signal effectively suppresses the hiss, which is otherwise present at the output of the superregenerator when it is not tuned to a radio station, or when the signal of such a station is too weak.

Under the influence of an incoming signal, the h.f. spurts are generated in such a receiver even when the tuned circuit of the receiver is adjusted to a frequency which considerably differs from the signal frequency. Of course, the amplitude of the signal will be lower in such a case, but as long as it exceeds the amplitude of the fluctuation pulses, the reception of the signal is still possible. Because of this effect, the selectivity of the superregenerative receiver is compar-

atively low, but the receiver possesses a much higher reception stability — with reference to a transmitter of unstable frequency — than the regenerative receiver.

The processes taking place in the superregenerator and discussed above explain the high sensitivity of receivers of this type. Even very weak incoming signals give rise to the local h.f. spurts in these receivers, the amplitude of such spurts increasing to a considerable value, the latter determining the audibility of the signals. The spurts are generated at the quenching frequency but the incoming signals govern the maximum amplitude of oscillations in each pulse, i.e., the signals control the generation-process of such oscillations.

The sensitivity of the superregenerative receiver depends, first of all, upon the value which can be reached by the amplitude of the locally-generated oscillations. When the operating condition of the receiver has been correctly set, the amplitude of the local signals can reach a value of several volts, although the incoming signal — causing the generation of the oscillations — often has a value of but some microvolts. Thus, the gain of the superregenerative receiver can be as high as several millions, and the value of the gain depends but slightly upon the amplification properties of the valve. It is possible to operate the superregenerator at a very low anode voltage (for instance at 15-20 v), if only the value of such voltage be sufficient to cause the self-excitation of the detector stage.

Analysing the processes taking place in the superregenerative receiver, it may be seen that, when the quenching voltage is generated by a separate oscillator, two types of operating conditions become possible, namely — the linear, and the non-linear (logarithmic) conditions. Under the linear condition, the oscillations being generated are given no time to build up to the maximum possible (set) amplitude. The amplitude of the oscillations is being constantly built up and reaches a certain peak value U_{max} at the moment when the self-oscillation is stopped by the increase of the bias voltage at the grid of the valve. This is the end of the oscillation and the beginning of its damping out. Curves given in Figs 233 and 234 correspond to just such a linear operating condition. This condition is obtained when the frequency of the quenching voltage is sufficiently high. Under the condition of linear operation, maximum amplitude U_{max} is proportional to voltage U_p of that original pulse which had caused the generation of the oscillations, i.e., it is proportional to the voltage of the incoming signal, and in the absence of the signal, it is proportional to the fluctuation voltage. If U_p is increased by some number of times, then U_{max} will also increase by the same number of times. Thus, a linear relation exists between U_{max} and U_p.

The superregenerative receiver, operating under the linear condition, introduces but very low distortion; but its gain in a great measure depends upon its supply voltages. If a stable amplification is to be obtained, these voltages must be stabilised. Pulse-type interference is poorly suppressed in a superregenerative receiver operating under the linear condition. Receivers designed to operate under such condition are difficult to adjust and are, therefore, seldom used by radio amateurs.

When the superregenerator operates under the non-linear condition, the amplitude of the generated oscillations is given sufficient time to reach a set value, i.e., the maximum possible value; and, once this value is reached, the amplitude remains constant during a certain period of time. In this case, U_{max} does not depend upon U_p. The initial voltage U_p influences only the building-up time of the oscillations. The greater the value of U_p, the shorter the period of time which will be required for the amplitude of the oscillations to reach the U_{max} value, and the longer will be the period during which the oscillations with constant amplitude U_{max} will take place. Under the condition of non-linear operation, the frequency of the quenching voltage must be lower than under the linear operating condition.

Fig. 235 gives a graphic representation of the oscillations on the reception of modulated signals by the superregenerative receiver operating under the non-linear condition. In the given case — as opposed to the case of the linear

operation — the amplitude changes of the incoming signal do not change the maximum amplitude, but change the duration of the h.f. spurts at the maximum amplitude of the generated oscillations. On detection, a certain average voltage value is obtained, as shown by the broken line in Fig. 235*d*. This voltage value is proportional to the duration of the spurts but is not proportional to the amplitude of the incoming signals. Hence, a considerable non-linear distortion is introduced, and this is the main shortcoming of the non-linear operating condition. However, this operating condition also offers its own advantages. First, it offers a stable amplification despite supply voltage changes. Secondly, it offers the feature of automatic gain control and the feature of attenuation of pulse-type interference. This is the reason why the non-linear operating condition is more frequently employed in superregenerative receivers.

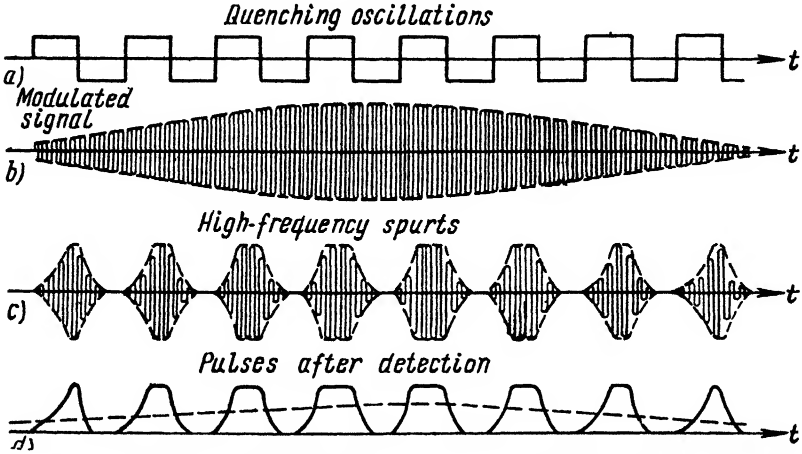


Fig. 235. A graphic representation of processes taking place during the reception of modulated signals in a superregenerative receiver working under non-linear operating condition

From the above discussion of the superregenerative receivers it follows that the quenching frequency must be, in all cases, in the supersonic frequency range. In other words, it must be beyond the range of audible frequencies but, at the same time, should be considerably lower than the frequency of incoming radio signals. If the last condition is not observed, the amplitude of the high-frequency oscillations does not reach a sufficiently high value during the positive half-waves of the quenching voltage. These requirements are hard to meet on the medium waves and even on short waves. On the ultra-short waves, the most advantageous quenching frequency values range between 100 and 200 kc.

It should be kept in mind that the superregenerative receiver radiates like a low-power radio transmitter, because it is kept in a state of oscillation. Because of this, it is highly desirable that the regenerative stage of such a receiver is preceded by a h.f. amplifier. The amplifier isolates the regenerative stage from the aerial, suppresses the radiation, increases the sensitivity of the receiver, and stabilises the operation of the set. If no such amplifier is provided in the superregenerator, then, apart from the very serious question of radiation, there arises yet another problem, namely, that of stability; any change of the aerial parameters in such a receiver will affect the tuning and the operating condition of the set, whose oscillatory circuit is directly coupled to the aerial. The h.f. amplifier may be omitted only in exceptional cases, for instance, when the superregen-

erator is designed for portable field service, when the number of valves and the power consumption of the receiver must be kept to the minimum.

Two general types of superregenerative receivers are encountered in practice. One of these types employs a self-quenching circuit, while the other one uses a separate quenching-frequency oscillator. An example of the receiving circuit incorporating a separate quenching-frequency oscillator is shown in Fig. 236. Here, valve V_1 is employed by the regenerative-detector stage as a capacitive-coupled ultra-short-wave oscillator. The capacitive coupling, in this case, is provided by the interelectrode capacitance within the valve. Tuned circuit L_1C_1 is adjusted to the frequency of the incoming signal. Valve V_2 is used in the quenching-frequency oscillator stage, in which the inductive type of feedback is provided. The quenching frequency is determined by the parameters of tuned circuit L_2C_2. The quenching frequency is fed to the superregenerative detector stage through capacitor C_3. Capacitor C_4 passes only the high-frequency currents

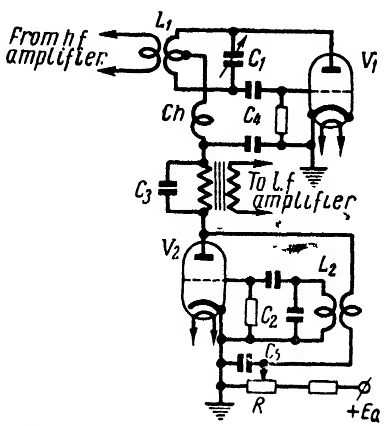


Fig. 236. The circuit diagram of a superregenerative detector employing a separate quenching-frequency oscillator

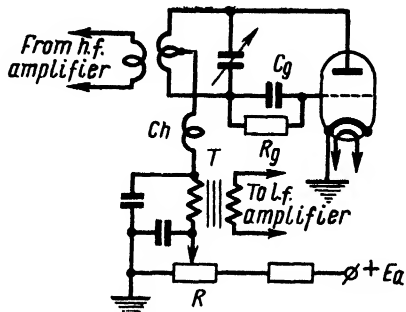


Fig. 237. The circuit diagram of a self-quenched superregenerative detector

of the signal being received, while the quenching-frequency currents, as well as the low-frequency currents, flow through capacitor C_5. Low-frequency signals, obtained as a result of detection, are fed by transformer T to a low-frequency amplifier. Choke Ch is used to block the high-frequency oscillations, keeping these oscillations away from the quenching-frequency and the low-frequency circuits. Potentiometer R is used to adjust the anode voltage supplied to the valves and serves to set the correct operating condition of the whole circuit. In this circuit, the oscillator employing valve V_2 functions as a modulator, interrupting the oscillations being generated by valve V_1 and its associated components. The circuit here described, as well as other circuits similar to it, are not favoured by the radio amateurs, because a separate quenching-frequency oscillator complicates the design of the receiver.

Superregenerators employing self-quenched circuits are considerably more popular than the type of receiver just described. In the self-quenched superregenerative detector, the periodic interruption of high-frequency oscillations is provided by a special operating condition of the grid circuit. A detector of such type is shown in Fig. 237. In this circuit, the negative bias, applied to the grid of the valve, is the voltage appearing across capacitor C_g, this voltage being the voltage drop built up across resistor R_g by the grid current flow.

Note, that in this circuit grid resistor R_g is connected not to the cathode, but to the positive anode voltage supply through the lower half of the tuned-

circuit coil. However, in some self-quenched detectors the grid resistor is connected to its usual point — to the cathode.

The operation of a self-quenched detector resembles the operation of a detector with a separate quenching-frequency oscillator working under the non-linear condition. On self-quenching, there is no linear relation between the low-frequency voltage, obtained as a result of detection, and the signal voltage. In other words, a considerable non-linear distortion occurs in this type of circuit. Here, the character of the high-frequency spurts is similar to the character of such spurts in a linearly-operated superregenerator, i.e., the amplitude of the oscillations reaches the maximum value, immediately after which the oscillations are damped out (Fig. 238).

As the amplitude of the oscillations increases, the grid voltage reaches the positive sector, thereby causing a grid current flow, which charges capacitor C_g. As a result, the bias voltage increases together with the amplitude of the oscillations, the bias increase being indicated by the thick broken line in Fig. 238. When the amplitude of the oscillations and the value of the bias voltage reach the maximum value, the condition necessary for self-excitation will cease to exist and the oscillations will begin to damp out. At the same time, the bias voltage will also decrease because capacitor C_g discharges through resistor R_g. But when the bias is decreased, the mutual conductance of the valve will again increase up to the operating point. Hence, at a certain moment the condition necessary for self-excitation will reoccur, and the incoming signal will make the detector generate the next h.f. spurt.

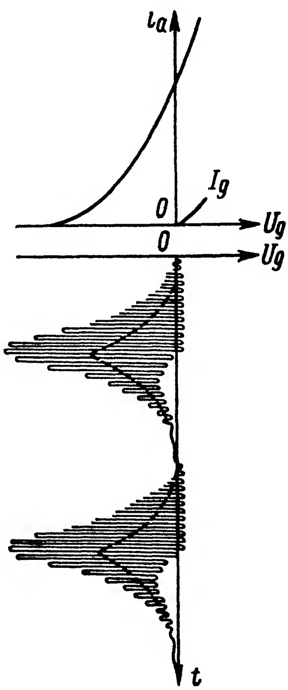


Fig. 238. A graphic representation of voltage at the valve grid in the self-quenched superregenerative detector

The greater the amplitude of the incoming signal, the more pronounced will be its action and the earlier will come the moment when the next bias-frequency h.f. spurt will occur. Thus, in contrast to the superregenerative receiver with a separate quenching-frequency oscillator, the self-quenched receiver is noted for the following peculiarity. The frequency at which its h.f. spurts appear is not constant, but is determined by the strength of the signals being received. At the same time, the duration of such spurts and the maximum amplitude of the oscillations remain constant in the self-quenched detector.

When there is no incoming signal, the frequency of repetition of the spurts in the self-quenched detector changes in a haphazard manner under the influence of the fluctuations. The detection will produce a certain voltage, this voltage also changing in a haphazard manner, although at a lower frequency than the repetition frequency of the spurts. The oscillatory process, in this case, will resemble the state of affairs depicted in Fig. 233. Now, however, the maximum amplitude of the spurts will remain constant, while their repetition frequency will be changing in a haphazard manner. As a result, the described action of the self-quenched superregenerator will result in the hissing, which is typical for such receivers.

A graphic representation of the processes taking place on the reception of modulated signals by the self-quenched superregenerative receiver is given in Fig. 239. As the amplitude of the incoming signal increases, the frequency at which the h.f. spurts reoccur also increases (Fig. 239*b*), and vice versa. However,

there is no direct proportion between the said amplitude and frequency. The average value of the rectified pulses, which are obtained after detection (this value is shown by the broken line in Fig. 239c), varies at the frequency of modulation. The higher the repetition frequency of the pulses, the greater will be the average value. However, the shape of the resultant low-frequency signal does not exactly correspond to the shape of the modulating signal, i.e., a non-linear distortion is present in the described case.

The optimum operating condition of the self-quenched receiver is set by changing the anode voltage with the help of potentiometer R. Varying the value of resistor R_g also helps to set the optimum operating condition. In the self-quenched regenerative receiver — just as it is observed in the case of non-linear operation of a receiver employing a separate quenching-frequency oscillator —

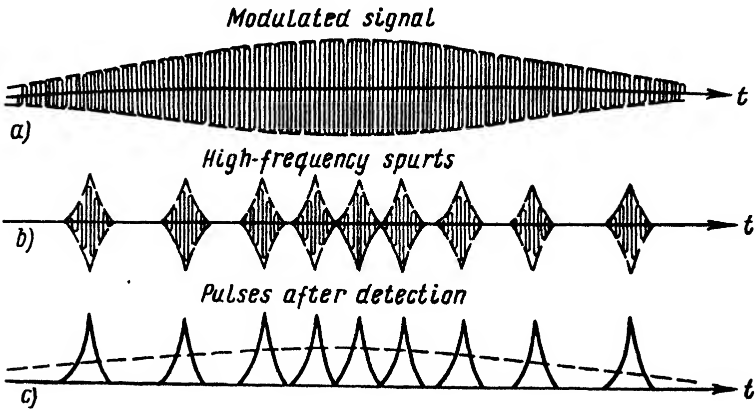


Fig. 239. A graphic representation of processes taking place during the reception of modulated signals in a self-quenched superregenerative detector

the superregenerator hiss and pulse-type interference are suppressed by the incoming signals. A good stability of operation and automatic gain control are also obtained.

Superregenerative receivers may be employed not only for the reception of amplitude-modulated signals, but also for the reception of signals carrying frequency modulation.

113. REFLEX RECEIVERS

The so-called reflex circuit is employed by some radio receivers. In such a circuit one and the same valve functions as a high-frequency amplifier and a low-frequency amplifier. This makes it possible to reduce the number of valves in the receiver and to cut down the consumption of the electric energy by the set. Both of these features are particularly important in battery-operated receivers, where the compactness of the set and long life of the batteries are at a premium.

Fig. 240 gives an example of the reflex receiver. This is a straight-amplification set, known as type 1-V-1, employing only two valves.

The incoming signal, picked up by the aerial, is passed on to the tuned input circuit L_1C_1, and then to the valve V_1. The amplified h.f. oscillations are developed in tuned circuit L_2C_2 and are fed to valve V_2, the latter functioning as a regenerative grid detector. The audio-frequency voltage, built up across resistor R_2, is then applied through isolating filter $C_4R_3C_5R_4$ and through gain control R_1 back to the control grid of valve V_1. Connected into the anode circuit of this valve is an electromagnetic loudspeaker LS (or the primary winding of an output transformer feeding a dynamic speaker).

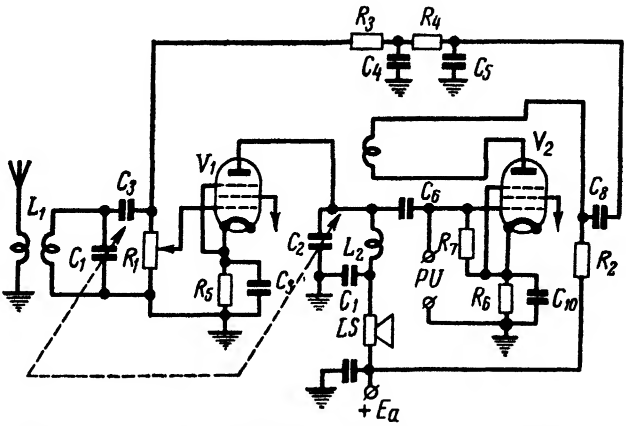


Fig. 240. The circuit diagram of reflex receiver, type 1-V-1

The function of the reflex circuit is based or the isolation (separation) of low-frequency and high-frequency oscillations. The capacitance of capacitors C_3, C_4, C_5 and C_6 is of the order of 200 pf, because these capacitors must pass only the high-frequency signals. Capacitor C_7, completing the path of high-frequency currents in tuned circuit L_2C_2 by earthing the bottom end of coil L_2, has a capacitance of a few thousand picofarads. (It must not pass the low-frequency current.) Isolating capacitor C_8, passing the audio-frequency component, has a value of 5,000-10,000 pf or higher. The resistance of gain control R_1 must be about 1 megohm. This potentiometer varies simultaneously the h.f. and l.f. gain. The resistance of the filter consisting of separate resistors R_3 and R_4 is about 0.1 megohm each.

When the radio receiver employing the described reflex circuit is called upon to reproduce gramophone records, the control grid of valve V_2 is automatically biased by the voltage drop developed across resistor R_6. For the sake of simplicity, the screen-grid circuits are not shown in the circuit diagram.

Reflex circuits may be incorporated into superheterodyne receivers, where a single valve can be used for the amplification of i.f.

and l.f. oscillations. It should be noted, however, that the reflex circuits give a somewhat lower performance than the usual circuits in which separate valves, working under different operating conditions, are employed for the h.f. and l.f. amplification.

114. VARIOUS METHODS OF CONTROLLING THE AMPLIFICATION, THE TONE AND THE SELECTIVITY OF RADIO RECEIVERS

Manual Gain Control

The basic purpose served by the manual gain control in a radio receiver is that of adjusting the level of the output audio signal to such a value which is most comfortable to the ear. Appropriate settings of this control also attain a certain amount of interference suppression (when the level of interference is lower than the level of the required radio signal). Finally, the manual gain control, when correctly set, prevents the overloading of receiver by excessively strong signals, which would otherwise cause distortion.

The manual gain control may be connected in various parts of a radio receiver. As shown under Chapter VII, and also under the present chapter (Sec. 110), the gain control may be connected at the input of l.f. amplifier, i.e., connected into the control-grid circuit of the first l.f. amplifier stage. Such method of connection offers a certain convenience because it allows us to control the volume of output sounds either when the receiver reproduces a radio broadcast or a gramophone record. This method, however, cannot prevent overloading of the h.f. and l.f. amplifier stages by excessively strong signals, because it offers no means of reducing the level of the signals passing through these stages. This shortcoming is overcome in the method of connection shown in Fig. 241, where the gain control shunts the aerial coil L_A. In this circuit, when the potentiometer slider is moved to decrease the value of R, the h.f. current flowing through the aerial coil is reduced. This reduces not only the level of the sounds reproduced by the loudspeaker, but also the signal input to all the stages of the receiver, thus guarding them against overloading.

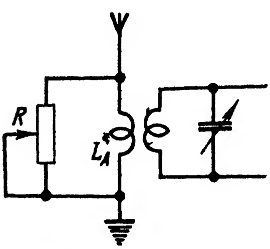


Fig. 241. The manual gain control incorporated in the input circuit of a radio receiver

Automatic Gain Control (A.G.C.)

The purpose served by the automatic gain control, usually referred to as a.g.c. and sometimes as automatic volume control or sensitivity control, is as follows. This control automatically equalises the amplification of the radio receiver when the level of the incoming radio signal varies as a result of fading. Such automatic equalisation prevents the overloading of the receiver and, at the same time, keeps the output signal level at a comparatively constant value in spite of fading of the h.f. signal level in the aerial.

The automatic gain control circuit utilises the d.c. voltage obtained as a result of signal detection. In radio receivers provided with the a.g.c. circuit, this voltage is fed as additional negative bias to the grids of the valves used in the stages preceding the 2nd detector. These valves are called *the controlled valves* and must possess remote cutoff characteristics to function properly with the a.g.c. circuit. If this is obtained, the amplification of the stages employing such valves will decrease with the increase of the supplementary negative bias voltage developed by the a.g.c. circuit. The stronger the incoming signal, the higher will be the d.c. voltage developed by the detector, the greater will be the negative bias voltage at the grids of the valves preceding the 2nd detector, and, hence, the lower will become the gain of the stages using these valves. In other words, a strong incoming signal automatically lowers the gain of the radio receiver, while a weak signal produces the opposite effect. As a result, the output signal of the receiver, incorporating the a.g.c. circuit, is maintained at an almost constant level.

As a rule, a.g.c. circuits are employed only in superheterodyne receivers. There are many varieties of a.g.c. circuits. Fig. 242*a* illustrates what is known as the *simple* a.g.c. circuit. The operating principle of this circuit is as follows. The controlled valves are initially biased with certain values of negative voltage, developed as a result of voltage-drops across self-biasing resistors R_c, connected into the cathode wires of these valves. The values of this initial bias are such that the valves operate over the parts of their characteristics where the mutual conductance is maximum. This provides for the highest gain in each stage. Besides the separate self-biasing resistors R_c, the detector load resistor R is included in the common grid circuit of the controlled valves. When a signal is picked up by the receiver, alternating audio-frequency voltage and d.c. voltage are built up across resistor R. The negative side of the d.c. voltage is applied to the grids of the controlled valves of the previous stages through the so-called a.g.c. filter (R_fC_f) and through decoupling grid filters $R_{f1}C_{f1}$ and $R_{f2}C_{f2}$. Thus, the grid circuit of each controlled stage is quite complex. In case of valve V_1, the grid circuit is comprised of the following parts, beginning with the chassis (the earth): cathode resistor R_{c1}, the space between the control grid and cathode

within the valve, coil L_1, the resistance of decoupling filter R_{f1}, the resistance of the a.g.c. filter R_f, detector load resistor R, and back to the chassis. In the absence of incoming signals, the grids of the controlled valves are biased only by the voltage developed across self-biasing resistors R_c. When the signals appear, the d.c. voltage developed across resistor R is added into the described grid-

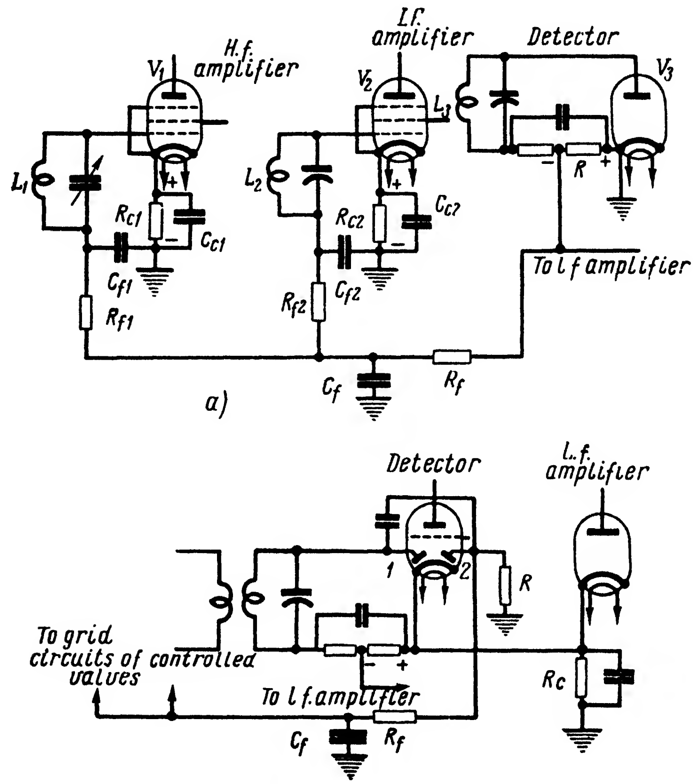


Fig. 242. A simple a.g.c. circuit *(a)*, and a delayed a.g.c. circuit *(b)*

circuits and increases the negative bias voltages applied to the grids of the controlled valves. The operating points on the characteristics of these valves are shifted by this additional voltage into the region of lower mutual conductance, thus reducing the amplification of the receiver.

The a.g.c. filter, consisting of resistor R_f (0.5-1 megohm) and capacitor C_f (0.05-0.1 mfd), serves to block the audio-frequency voltage appearing across R, preventing this voltage from reaching the grids of the controlled valves. Because the d.c. voltage built up across the same resistor R changes comparatively slowly as the

signal fades or builds up, this changing d.c. voltage is easily passed through R_f to the grids of the controlled valves. In comparison with this, R_f offers a much higher resistance to the audio-component than does the capacitive reactance of C_f. As a result, the low-frequency voltage is almost completely dropped in R_f and only a very small part of such voltage appears across capacitor C_f. This voltage reaches the grids of the controlled valves but does not upset the normal operation of the valves. Thus, filter R_fC_f acts in a manner similar to the usual decoupling filter. A separate decoupling filter, connected into the grid circuit of each controlled valve, eliminates the parasitic coupling that could otherwise be established between the controlled stages.

The shortcoming of the simple a.g.c. circuit just described is that this type of circuit begins to operate even when the level of the incoming radio signal is low, which prematurely decreases the amplification of the receiver. Because of this, such a.g.c. circuits are seldom used, preference being given to the so-called *delayed* a.g.c. circuit. In the delayed a.g.c. circuit, the reduction of the amplification begins only when the signal strength reaches a certain predetermined value and begins to exceed it. In such a circuit, there is no further weakening of amplification of those incoming signals which are already weak enough. The circuit, however, requires an additional diode, as explained below.

One of several versions of the delayed a.g.c. circuit is shown in Fig. 242*b*. The circuit employs a double-diode valve, in which diode *1* functions as the ordinary diode detector with the usual load resistor, across which the audio component is built up and fed to a l.f. amplifier. Diode *2* of the same valve functions as the a.g.c. detector. The signal is applied to diode *2* through capacitor C, while the voltage rectified by this diode appears across its load resistor R. Cathode resistor R_c of the l.f. amplifier stage is connected into the circuit of diode *2* and, because of this, a constant negative bias voltage is applied to the anode of the a.g.c. diode, i.e., to the anode of diode *2*. This bias voltage is called the delay voltage and its value is determined by the voltage drop across R_c.

Let us assume that the delay voltage is equal, for example, to 3 volts. Then. as long as the signal voltage does not exceed 3 volts, the a.g.c. diode does not pass current and the a.g.c. circuit does not function. Under this condition, the amplification of the stages preceding the detector is at its maximum value. As soon as the amplitude value of the signal begins to exceed 3 v, the a.g.c. diode begins to conduct current, this current producing a voltage drop across load resistor R. The d.c. component of this voltage will be fed through filter R_fC_f to the grids of the controlled valves, reducing their amplification.

The d. c. voltage developed by the a.g.c. detector is sometimes insufficient to effect the amplification control in the stages preceding

the detector. If such be the case, the so-called *amplified a.g.c. circuit* must be used to intensify the influence of the automatic gain control upon the controlled stages. The amplified a.g.c. circuit employs an auxiliary i.f. amplification stage, this stage feeding amplified voltage to the a.g.c. detector. This makes the detector develop a sufficient d.c. voltage, thereby intensifying the action of the a.g.c. circuit upon the controlled stages.

It should be kept in mind that the automatic gain control cannot completely counteract the fading of radio signals. If the signal level falls to a negligible value, the a.g.c. circuit, of course, cannot restore the normal loudness of sound reproduction.

Tone Control

Most radio receivers designed for the reception of broadcast programmes incorporate a tone control. The tone control circuit is a part of the l.f. amplifier circuit and it provides a means of changing the pitch of the reproduced sounds. Various tone control circuits were described in Sec. 78, Chapter VII.

Selectivity Control

Modern superheterodyne receivers sometimes employ special selectivity-varying circuits. Such circuits are known as *selectivity control circuits*. The selectivity control circuit is a circuit which provides a means of controlling the frequency bandwidth passed by the receiver, thus controlling the degree of selectivity. In the presence of interference to the radio reception, the selectivity control circuit can be adjusted to increase the selectivity of the set. This cuts off a considerable part of interference by narrowing the frequency bandwidth passed by the receiver. But this impairs also the fidelity of reproduction. When the reception is not being interfered with, the selectivity control is adjusted to decrease the selectivity of the radio receiver. This broadens the frequency bandwidth being passed and provides for high-fidelity reception.

Practical selectivity-control circuits utilise the effect of varying the coupling between the tuned circuits of the i.f. transformers. The coupling may be varied in steps or continuously. The continuous adjustment of coupling is attained by varying the mutual inductance between the tuned-circuit coils in the i.f. transformers, as shown in Fig. 243*a*. The step adjustment of coupling is effected by connecting auxiliary coupling coils between the tuned circuits (Fig. 243*b*). The closer the coupling, the broader will be the frequency band passed by the receiver.

Another method of selectivity control consists in detuning the tuned circuits of i.f. stages. An alternative method is that of connecting variable ohmic resistances into such tuned circuits. In some cases, the width of the frequency band is varied by means of special low-frequency filters connected into the low-frequency amplifier circuit of a radio receiver.

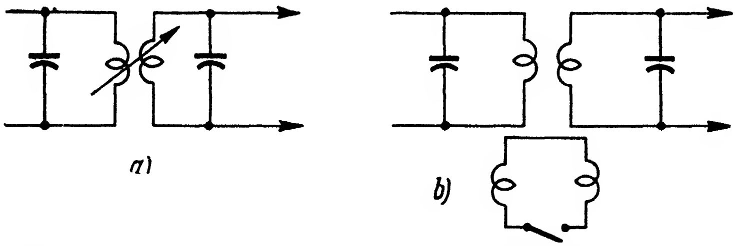


Fig. 243. The adjustment of selectivity in an intermediate-frequency amplifier

The simplest types of straight-amplification receivers attain the control of selectivity by varying the coupling between the aerial and the input circuit of the receiver. In such circuits, the closer the coupling, the lower the selectivity, and, conversely, the looser the coupling, the higher the selectivity.

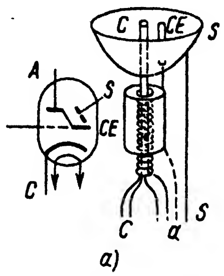
115. THE "MAGIC EYE"

Modern superheterodyne receivers make wide use of a special tuning indicator, known as the "magic eye". When the "magic eye" is provided in the receiver circuit, the receiver can be tuned to exact resonance with the frequency of the incoming signal, even when the manual volume control is set to the position of zero sound. Once the receiver is tuned to station in the described way, it is only necessary to advance the gain control knob to such a position where the station will come in at a comfortable sound level. Such "silent" tuning offers apparent advantages. Hence — the popularity of the "magic eye" with radio listeners, although the device is also used in certain types of measuring apparatus.

The design principle of the "magic eye" and its schematic representation are given in Fig. 244*a*. The device has the shape of an ordinary glass valve, in whose envelope are located a triode and an electron-beam indicator, this indicator being the heart of the "magic eye". In this valve, cathode *C* performs its usual function of electron emission. The cone-shaped screen *S* is the anode of the valve. The inner part of this screen is covered with a luminescent material (willemite), which glows green under the impact

of electrons. The third electrode of this valve is the control electrode, designated as CE in the drawing. This electrode has the shape of a narrow strip and is connected to the anode of the triode section.

Such is the general design conception of the "magic eye". Now, let us see how the device is connected with its external circuit (Fig. 244*b*). The grid of the triode section of the "magic eye" is supplied with voltage from the load resistor of the detector. The anode of this triode section is connected with screen S of the indicator through resistor R_a (1-1.5 megohms). When the receiver incorporating such a circuit is not tuned to any radio station, the anode current of the said triode section builds up a certain voltage drop across resistor R_a, the negative potential being applied to control electrode CE. Thus, the control electrode is charged negatively in respect to screen S. Because of this, the electric field surrounding the control electrode is such, that the



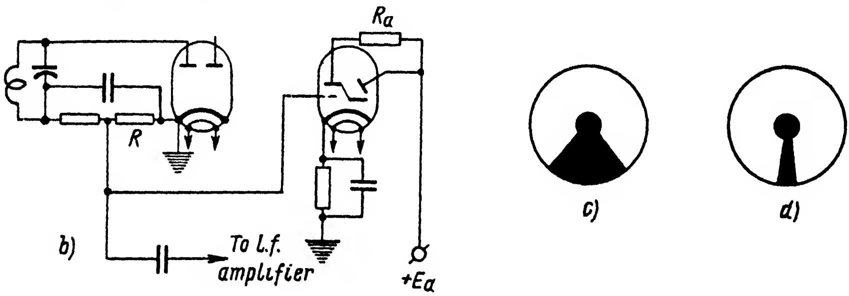


Fig. 244. The circuit diagram, mechanical design, and action of the "magic eye".

path of the electrons given off by the cathode is curved in the vicinity of CE, the electrons being attracted to that part of S which is free of the shielding effect of CE. As a result of the above action, the inner surface of screen S will glow, with the exception of the dark sector formed in the region of control electrode CE. This dark sector is commonly referred to as "the shadow". The shape of this shadow in the absence of signals is shown in Fig. 244*c*.

When the radio receiver is tuned to the incoming signal, the negative voltage built up across the detector load resistor R is applied to the control grid—which is different from the control electrode mentioned above—of the triode section in the "magic eye" valve. This will cause a reduction of the anode current in the

said triode section, thereby decreasing the voltage drop across resistor R_a. As a result, the potential of control electrode CE will become more positive, the value of the positive potential of this electrode approaching the potential value of screen S. Owing to this, the deviation of the electrons from their straight-line path will become smaller, thus accounting for the narrowing of the shadow on the screen (Fig. 244*d*). When the radio receiver is tuned to exact resonance with the frequency of the incoming signal, the shadow will assume its narrowest shape. If the strength of the incoming signal is sufficiently large, the anode current of the triode section of the "magic eye" valve will be almost completely cut off. In such a case, the potential of control electrode CE will be nearly the same as the potential of screen S, because the voltage drop across resistor R_a will be negligibly small. This will cause the shadow to disappear completely, and the whole of screen S will now glow.

As a rule, the triode section of the "magic eye" is not included in the main amplification channel of the radio receiver. Because of this, the receiver will function normally with or without the "magic eye". There are, however, certain superheterodyne circuits in which the triode section of the "magic eye" is a part of the circuit of the first low-frequency amplifier stage. It is also possible to design a "magic eye" in which the triode section is replaced by a pentode. The Soviet valve-manufacturing industry produces two types of "magic eyes"—type 6E5C and type 6E1П. Both types of valves are provided with triode sections.

116. INTERFERENCE AND METHODS OF ITS ELIMINATION

Atmospheric Disturbances

Atmospheric disturbances, sometimes referred to simply as "atmospherics", are the electromagnetic waves produced by various types of electric discharges in the atmosphere. The duration of such discharges does not exceed a few thousandths of a second, and they are always taking place in the earth's atmosphere. The intensity of discharges reaches its maximum in summer, while in winter it falls to such a low value that it practically ceases to interfere with radio reception.

The atmospheric disturbances are reproduced by the loudspeaker of a radio receiver as crashing and snapping noises, rustles and hisses. This type of interference is considerably less noticeable on short waves (10-50 m) than on the long waves, and is practically absent in the ultra-short-wave range. The atmospheric disturbances may be classified as local and distant. The local disturbances are caused by thunderstorms and by silent discharges in the locality of

the receiver. The distant disturbances come from far-away lands, for instance, from Central Asia and Africa, where the centres of continuous atmospheric disturbances are located.

Industrial Noise

The type of interference created by various electrical machinery and devices is known as industrial noise. Another and quite appropriate name for the given type of electrical noise is "man-made noise". This noise is practically absent in localities situated far away from different concentrations of electrical machinery. In cities, however, the industrial noise can at times reach such a high level as to make radio reception impossible. As a matter of fact, this type of noise aggravates the radio reception to a much greater extent than the atmospherics.

Industrial noise is mainly caused by spark discharges in different electrical apparatus and devices. The chief offenders in the cities are the spark discharges taking place between the current-carrying conductors and the trolleys of trolleybuses and tramcars. These discharges are reproduced by radio receivers as crashing and snapping noises and rustles. The electrical noise generated by automobile ignition systems is reproduced as a continuous crashing noise, heard on the long-wave, short-wave and even ultra-short-wave bands of a radio receiver.

Various electrical medical devices, X-ray apparatus, arc-type motion-picture projectors, mercury-arc rectifiers and arc welders also generate continuous crashing noise in the loudspeaker of a radio set, such noise frequently blanketing the reception over a wide frequency range. It is easy to determine when the offending apparatus is switched on and off, for this is immediately accompanied by the respective commencement and cessation of the described type of noise at the output of radio sets.

Man-made noise is also produced by sparking in various switches, in electrical bells, between the collectors and brushes of electrical motors and generators, in the automatic switching systems of electrical advertisements, traffic lights, etc.

As follows from the above, any device in which electric sparking is developed may be a source of industrial noise. The sparking induces high-frequency currents in various wires. These currents flow through the wires and radiate electromagnetic waves into the surrounding space, thus generating the man-made noise interfering with radio reception. Such noise can reach the radio receiver circuits in any one of the three following ways: 1) through the receiving aerial; 2) through the power mains, if the receiver operates from such mains; 3) by induction from the mains or other wires along which the offending noise signals propagate.

Receiver Noise

Quite independently of the atmospherics and man-made noise, a radio receiver — particularly if it is of a high-gain type — can develop its own internal noise. Such noise will be reproduced by the loudspeaker of the set and will interfere with the radio reception, although not as badly as the external electrical noises discussed above. The internal noise of a radio receiver is generally referred to simply as *receiver noise*. The receiver noise is generated by various irregularities of the thermal emission within electron valves. (This effect is known as the shot effect. The receiver noise is also attributed to the haphazard movement of electrons in various resistors and in the wiring of the radio receiver — this effect known as the fluctuation noise.—*Translator's note.)*

Receiver noise is almost absent at the output of simple straight-amplification sets, but it can be quite strong and very unpleasant at the output of a superheterodyne receiver employing a large number of stages and possessing a high gain.

The a.c. hum produced in mains-operated radio sets should also be included under the heading of receiver noise. If a radio receiver is properly designed and its rectifier is provided with a sufficient degree of filtering, the a.c. hum will be hardly heard in the loudspeaker. The presence of hum at the output of a radio receiver is an indication that something is wrong with the set.

Various Methods of Noise Elimination

Noise produced at the output of a radio receiver by atmospherics is most difficult to eliminate. Such noise is usually weak in straight-amplification sets employing no regeneration. The incorporation of regeneration in a receiver increases the sensitivity of the set, but it also raises the noise level. Superheterodyne receivers are most sensitive to the noise produced by the atmospherics. In these receivers the atmospheric noise may be somewhat reduced by increasing the selectivity of the set. Another way to reduce this type of noise is to operate the receiver from a loop aerial. This, however, is effective only when the direction to the required radio station does not coincide with the direction in which the atmospheric disturbances are generated. In order to reduce the atmospheric interference as much as possible, it is recommended to use short and low-slung receiving aerials, i.e., to reduce the natural capacitance of the aerial.

The above methods of suppression of the atmospherics could also be applied to the suppression of the man-made noise. This, however, usually does not give the desired effect, because the industrial noise frequently reaches the receiver circuit not via

the aerial but by other means. Hence, when dealing with the man-made noise, other methods of suppression have to be devised. Thus, in a number of cases the man-made noise can be suppressed right at its source. For instance, the electrical noise generated by tramcars can be almost totally suppressed if the usual aluminium trolley runner is replaced by a carbon one. Metal shielding of the source of interference is a good way of suppressing the electrical noise. As an example — this type of noise suppression is employed in

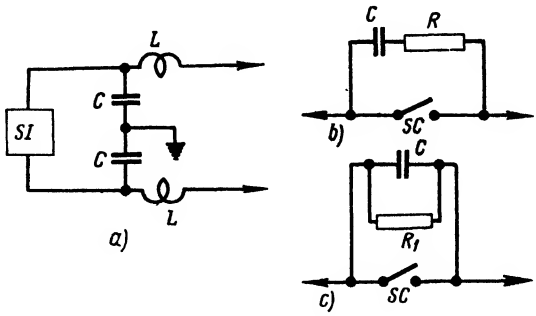


Fig. 245. Filters suppressing the electrical noise generated by various sources of industrial interference *(SI)* and by sparking contacts *(SC)*

automobiles, where all parts of the ignition system are shielded. (In a car, only the spark plug and the distributor wires can be shielded. The shielding has to be supplemented by earthing various components of the ignition system — in respect to high-frequency — through capacitors. Even this is insufficient, and the complete noise suppression is obtained only when high-value resistors are connected in series with the spark plug leads.—*Translator's note.*)

High-frequency filters, comprising chokes and capacitors and connected into the wires running from the offending electrical device, provide a simpler method of noise suppression. When such a method is employed, high-frequency currents will be prevented from flowing along the wires, i.e., prevented from spreading interference via induction and radiation. Of course, it is not always possible to incorporate such filters into the circuit of every electrical device. Still, they should be used wherever possible, since they present a simpler way of noise suppression than does the shielding. The most effective noise suppression method consists in employing both the filters and shielding. However, in this method the filters are the main thing, because they stop the interference from propagating along various conductors, via which the interference usually reaches radio receivers.

Fig. 245*a* gives the circuit diagram of a filter designed to prevent the penetration of interference into the wires running from the

source of such interference. The source, in the given case, is denoted by letters *SI* in the drawing and may be represented by an electric motor, electrical welder, electrical medical device, switch, bell, etc. The filter consists of capacitors and high-frequency chokes. The chokes block the high-frequency currents which otherwise would be propagated from *SI* along the wire line, while the capacitors, possessing low capacitive reactance, bypass these currents to earth. The metal frame of interference source *SI* should also be earthed. The chokes must have an inductance of about 1 mh, their distributed capacitance should be made as low as possible, and the wire with which they are wound should be computed to carry the rated current. The capacitors should have a capacitance of approximately 0.1 mfd and must be rated at correct working voltage. The interference developed as a result of sparking between various types of contacts (sparking contacts *SC)* should be suppressed with the help of circuits shown in Fig. 245 *b* and *c*. Here, the ohmic resistors must have the following values: $R = 50\text{-}100$ ohms, $R_1 = 5{,}000\text{-}100{,}000$ ohms. The value of the capacitors shown in these circuits should be 0.1 mfd.

In order to reduce the pick-up of interference by the aerial of a radio receiver, the aerial should be installed as far as possible from various sources of electrical noise, for instance — from power and light mains, from telegraph lines, signalling lines, etc. It is particularly important that the aerial is located far away from the wiring belonging to tramcar and trolleybus electrical systems.

It is highly desirable that the aerial of a radio set is made perpendicular to the offending electrical line. If this is done, the interference propagated by tramcar lines and other conductors becomes negligible at distances of about 100 metres from the aerial. In order to prevent interference pick-up by the aerial downlead, the latter should be shielded. Special downlead wire, in which earthed shielding braiding is placed over the wire insulation, is available. When the wire of this type is employed to connect a radio receiver with its aerial, the shielding braiding must cover all parts of the downlead from the point of its connection to the aerial to terminal marked "Aerial" on the radio set. This will preclude the possibility of interference reaching the downlead wire from nearby lighting leads in the room. Fig. 246*a* shows how the aerial should be connected to the shielded downlead. Incidentally, if the special downlead of this type cannot be readily purchased, the receiver owner can devise a "home-made" variety of the required shielded wire. It is only necessary to take any ordinary type of rubber-insulated wire and to wind a wire spiral over the insulation. The spiral may be wound with 0.5-0.8 mm bare or insulated copper wire. In such a spiral winding over the downlead wire the adjacent turns may be as far from each other as 0.5-1 cm. Be sure to earth the spiral. Keep in mind, that the shielding placed over the downlead considerably lowers the h.f. voltage fed by the

aerial to its receiver. Consequently, only highly-sensitive receivers should be provided with such downleads, and then only when bad interference is experienced in the given locality.

Fig. 246*b* shows the design of the so-called "noise-suppressing aerial". The part of this aerial comprising the coil L_1 is the part which does the actual picking up of radio signals. The given part should be installed on the roof of a building, as far away as possible from all sources of interference. This part of the aerial may be represented by a short or slanted piece of wire. The shorter the wire, the better, because it will pick up less interference. A weatherproof h.f. transformer L_1L_2 is built-in right into this piece of wire in such a way that one end of coil L_1 is connected to the earthing lead, while the other end is connected to the pick-up part of the aerial wire. Coil L_2 of the given h.f. transformer is connected to the remote radio receiver by means of a two-wire line, known as the feeder. The feeder runs from the transformer right up to the radio set, where it terminates in another h.f. transformer L_3L_4. A special earthed static screen is placed between the primary and secondary windings of transformer L_3L_4. This screen is a mesh of thin insulated wires connected to a single point, the latter point, in its turn, being connected to earth. The described screen acts to eliminate the capacitive coupling between the two coils of h.f. transformer L_3L_4.

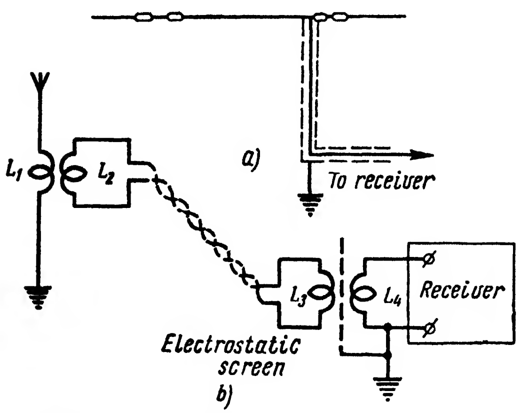


Fig. 246. The aerial with a shielded downlead *(a)*; and the "noise-suppressing" aerial with a two-wire feeder *(b)*

The "noise-suppressing aerial" functions in the following way. Radio signals received by the pick-up part of the aerial are directed to the receiver over the feeder. The signals are slightly attenuated in the feeder. However, this loss of the useful h.f. energy is more than compensated for by nearly complete suppression of interference that might be picked up by the feeder. The interference, in this case, is suppressed because the e.m.f. of the interfering electrical noise sets up h.f. currents in both conductors of the feeder. These currents are of equal value and flow in the same direction so far as feeders are concerned, but in the opposite directions so far as coils L_2, L_3 are concerned. Thus in the transformers their effect is cancelled and they are kept away from the input circuit of the radio receiver.

It is, of course, impossible to get rid of interference completely because the acting part of the described aerial will pick up some interference and will convey it to the receiver, no matter how high this aerial part is strung up over the surface of the earth.

When a radio receiver is made to operate from an indoor aerial, a strong industrial noise is often experienced, because such an aerial is generally located close to house lighting-wiring, and, at the same time, is not very efficient at picking up radio signals. As a result, the ratio of useful signal strength to the strength of the noise voltage is not sufficiently high for good reception.

There are several ways of preventing the industrial noise carried by light mains from reaching the circuits of radio receivers powered from such mains. One way is that of placing a screen between the primary winding and the secondary windings of the power transformer of the receiver. The screen can be made of a thin metal sheet, but such a sheet must not form a short-circuited turn. Alternatively, a special screening winding can be placed between the transformer windings. The screening winding may be wound in one layer with 0.15-0.2 mm wire and should be connected to earth (to the common negative terminal of the receiver). Another way of isolating the receiver from the electrical noise carried by the mains is the incorporation of an electrical filter into the mains wires running to the receiver. Such a filter is quite similar to the h.f. filters connected into the wiring running from interference sources (Fig. 245*a*). The inductance of the filter should be about 0.5-1 mh, while the capacitance value of the capacitors it employs must be approximately 0.1-2 mfd.

117. THE RECEPTION OF FREQUENCY-MODULATED SIGNALS

The rapid growth of frequency-modulation broadcasting (usually referred to simply as FM broadcasting) makes it mandatory that every student of radio is well versed in the principles and operation of radio receivers designed for the reception of frequency-modulated signals. Fig. 247*a* gives the block diagram of a FM receiver, whose general functions should be thoroughly understood before a discussion on its individual circuits can be undertaken.

The FM receiver is nothing but a superheterodyne set. This receiver, however, differs from the usual superhet, employed for the reception of amplitude-modulated signals. The points of difference are seen in a specific operating condition of the 2nd detector and in the employment of a special limiter stage preceding such detector. As we already know, a frequency-modulated signal is noted for the constancy of its amplitude — the modulation applied to such a signal varying not the amplitude but the frequency of the carrier wave. It should be noted, however, that various types of interference, encountered in the transmission of the signal from the radio station to the receiver, account for certain amplitude changes of a frequency-modulated signal. By the time such signal reaches the detector stage of the radio receiver, the signal not only carries the normal frequency modulation, but also carries a certain amount of undesirable amplitude modulation, caused by the interference.

It is the purpose of the special limiter stage, employed by every FM receiver, to decrease the unrequired amplitude modulation named above, thus lessening the interference at the output of the receiver.

Fig. 247*b* gives the characteristic of the limiter and shows the dependence of its output voltage U_0 upon input voltage U_i. When the value of U_i is small, voltage U_0 increases proportionally to voltage U_i. However, when the input voltage reaches a certain value, its further increase does not cause any further increase of U_0. This value, at which the limiting action begins, is known as the threshold of limiting. Fig. 247*b* gives the curves of input voltage U_i, carrying noise modulation, and of the corresponding output voltage freed of such modulation.

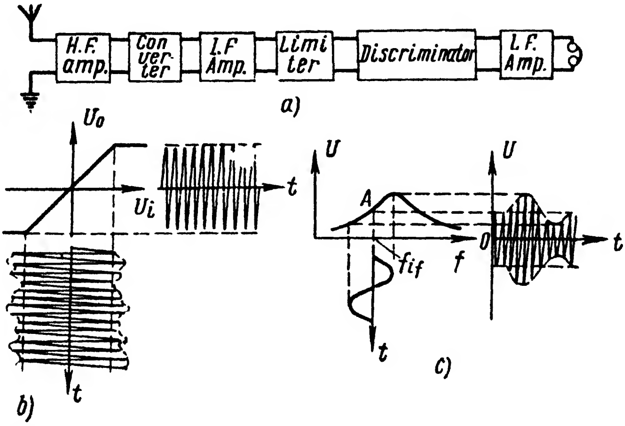


Fig. 247. The block diagram of an FM receiver *(a)*. Graphic representation of the operation of the limiter *(b)* and of the discriminator *(c)*

The limiter stage can employ various circuit arrangements. For example, an intermediate-frequency amplifier stage of the receiver can act as the limiter, if the anode and screen-grid voltages of such a stage are reduced to about +10 volts by means of high resistances connected into the anode and screen-grid circuits. In the limiter stage of this type, a high resistance must also be connected into the control grid circuit. When this is done, the limiting action will be obtained by virtue of grid currents, which sharply increase on the reception of stronger signals.

The detection of frequency-modulated signals can, similarly, be performed in various ways. The simplest type of frequency detector employs a tuned circuit which is slightly out of resonance with the frequency of the signal. The operation of such tuned circuit takes place along the drooping side of the resonance curve (Fig. 247*c*). Point A on the resonance curve corresponds to the intermediate frequency f_{if}. The curve plotted below the resonance curve shows the frequency changes of the oscillations fed to the detector. These frequency changes are subsequently converted into amplitude changes, as shown in the graphic representation at the extreme right. Thus, the voltage built up across the tuned circuit is modulated in amplitude, the amplitude changes approximately corresponding to the changes of frequency. These amplitude-modulated oscillations are then fed to the usual type of detector. This detector can employ any suitable circuit, audio-frequency oscillations developing at its output.

Speaking again about the detection of the frequency-modulated signals, it should be noted that the best results are obtained by employing detectors specifically designed for this type of duty and known as discriminators. The

circuit of a simple discriminator is shown in Fig. 248*a*. In this circuit, tuned circuits LC and L_1C_1 are adjusted to intermediate frequency f_{if}, this frequency carrying the converted h.f. signal. Note that coil L_1 is centre-tapped. Because of the centre tap, voltages U_1 and U_2 are equal in amplitude but are shifted in phase by 180°. Diodes *1* and *2* are usually represented by crystal- or vacuum-type diodes (in the latter case, a double-diode electron valve with separate cathodes may be employed). Resistors R_1 and R_2 are the load resistors. The capacitive reactance of capacitors C_2, C_3 and C_4 is low, as far as the high-frequency currents are concerned. On the other hand, choke Ch presents a high inductive reactance to currents with frequency f_{if}. When the input circuit of the discriminator is fed with frequency-modulated voltage U from the limiter, audio-frequency voltage U_0 is developed at the output of the discriminator. Let us see how this voltage is produced.

As seen from the circuit diagram, voltage $U + U_1$ acts in the circuit of detector *1*, while voltage $U + U_2$ acts in the circuit of detector *2*. In order to determine the value of these voltages, we have to know the phase shift between them.

If the frequency of the input voltage is equal to resonant frequency f_{res}, then the voltage across the secondary tuned circuit is shifted by 90° in respect to voltage U. This can be shown to be true as follows.

The current flowing through L lags behind voltage U by 90°. The higher the rate of change of this primary current, the greater will be the value of e.m.f. induced in the secondary coil. But the highest rate of change is obtained when the current passes through its zero value. It is at this moment that the induced e.m.f. reaches its highest value. Therefore the e.m.f. of the secondary tuned circuit lags behind the current in the primary by 90°, and lags behind voltage U by 180°. Tuned circuit L_1C_1 is adjusted to resonance. The current flowing through it is in phase coincidence with the e.m.f., i.e., this current lags behind voltage U by 180°, while the voltage across coil L_1 leads the current in the secondary circuit by 90°. Hence, there is a 90° phase shift between the primary and secondary voltages. Such phase shift is typical of all inductively-coupled tuned resonant circuits.

In the circuit being described, the secondary voltage is divided into two voltages U_1 and U_2, these voltages being of equal amplitude but of opposite phase. Because of this, if U_1 lags behind U by 90° (a phase shift of +90°), then U_2 leads voltage U also by 90° (a phase shift of —90°). It now becomes possible to determine the values of voltages $U + U_1$ and $U + U_2$. This is done in Fig. 248*c*, where, as an example, the amplitudes of voltages U, U_1 and U_2 are made equal. As may be seen, the amplitudes of both resultant voltages are equal. When these voltages are rectified by detectors *1* and *2*, d.c. voltages will be built up across resistors R_1 and R_2 and the given d.c. voltages will be equal to each other, although they will be opposite in polarity. As a result, the output voltage of the discriminator will be zero in the described case.

Continuing our study of the discriminator action, let us now take the case when the frequency of the input voltage changes, for instance — decreases. The tuned circuit will, thus, be out of resonance with the frequency of the signal. The capacitive reactance will predominate in this tuned circuit and the current will lead the e, m.f. Let us assume that the phase angle of such lead is 30°. This means that the secondary current now lags behind the primary voltage by 180°—30° = 150°. The voltage across the secondary coil L_1 leads the current by 90° and, hence, this voltage now lags behind voltage U by 150°—90° = 60°. Since voltage U_1 and U_2 are always in phase opposition, the phase shift between U and U_1 will be equal to +60°, while the phase shift between U and U_2 will be equal to —120°. This case is illustrated by the addition of voltages in Fig. 248*c*. As may be seen from this graphic representation, the resultant voltage $U + U_1$ has increased, while the resultant voltage $U + U_2$ has decreased. Because of this, the rectified voltage across R_1 will be greater than the rectified voltage across R_2. Therefore, the output voltage, determined by the difference of the voltages across R_1 and R_2, is no longer equal to zero and has a definite positive value in respect to earth.

In the above discussion, we had assumed that the frequency of the signal is changed in such a direction, that this frequency becomes lower than the frequency to which the tuned input-circuit of the discriminator is adjusted. This, as we have seen, had resulted in building up a positive resultant d.c. potential across the discriminator load resistors R_1 and R_2. Going through the same discourse, but taking the opposite case — i.e., when the frequency of the signal swings into such a direction that this frequency becomes higher than the fre-

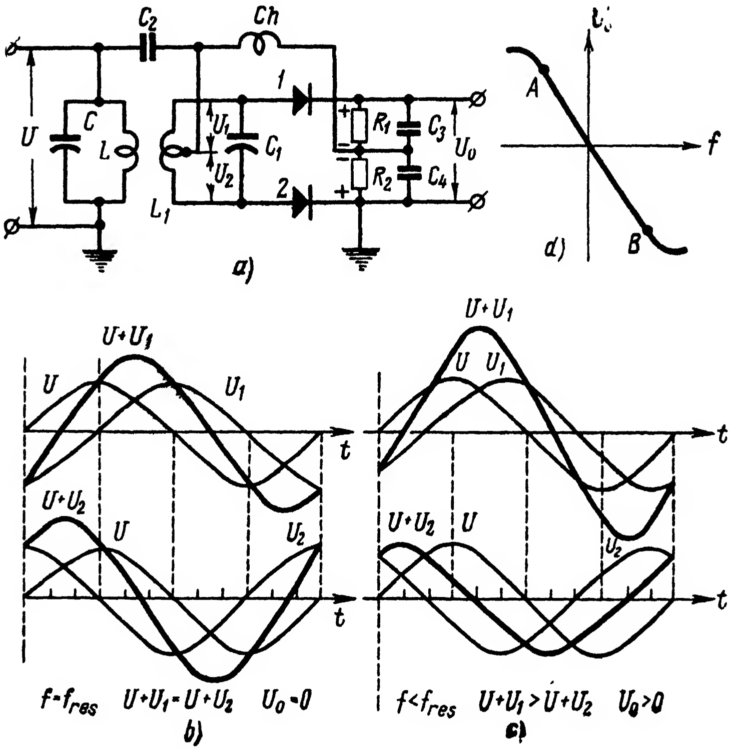


Fig. 248. The circuit diagram of a discriminator *(a)*. Voltage curves *(b* and *c)*, illustrating the operation of the discriminator. The discriminator characteristic *(d)*

quency to which the tuned input-circuit of the discriminator is adjusted — we find that $U+U_1$ becomes smaller than $U+U_2$, and, therefore, a negative resultant d.c. potential will be built up across resistors R_1 and R_2 in such a case.

In both of the above cases, the greater the deviation of the signal frequency from the frequency to which the tuned input circuit of the discriminator is adjusted, the greater will be the voltage developed across the discriminator load resistors. This, however, can be carried on only up to a certain limit. This limit will be reached when the tuned circuits are so much detuned in respect to the signal frequency that the voltage built up across them is lowered as a result of such detuning. When the described limit is reached, the output voltage of the discriminator will drop. The dependence of output voltage U_0 upon the signal frequency is illustrated by the discriminator characteristic given in Fig. 248*d*. Under normal conditions, the frequency discriminator operates over the linear part of *AB* curve and provides an undistorted conversion of

frequency-modulated oscillations into audio-frequency signals. (In some countries, this type of discriminator is known as Foster-Seeley discriminator. — *Translator's note.*)

In certain versions of discriminator circuits, the discriminator simultaneously acts as a limiter. A circuit possessing this property and known as the "phase-relationship" discriminator is given in Fig. 249.

The phase-relationship discriminator differs from the discriminator shown in Fig. 248*a* as follows: in the phase-relationship discriminator diodes D_1 and D_2 are connected in series, while their load resistors R_1 and R_2 are shunted by high-capacitance capacitors C_5 and C_6, the value of these capacitors being about

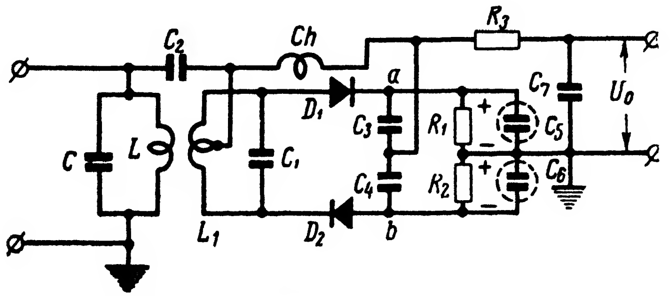


Fig. 249. The circuit diagram of the "phase-relationship" discriminator

10 mfd. Besides, the given type of discriminator employs capacitors C_3 and C_4, the value of such capacitors being sufficiently small to oppose a flow of audio-frequency currents. In the phase-relationship discriminator, the output voltage is obtained between the earth and the point where capacitors C_3 and C_4 are connected to each other. This output voltage is then fed to a low-frequency amplifier through filter consisting of resistor R_3 and capacitor C_7.

When the detector of this type is in operation, the voltages built up across capacitors C_5 and C_6 do not change at audio frequency because of the high capacitance value of these capacitors. Likewise, the voltage across points *a* and *b* is also constant, this voltage being the resultant voltage built up across capacitors C_5 and C_6. If the signal carries a certain amount of undesirable (parasitic) amplitude modulation, such modulation will not affect the value of voltages normally built up across capacitors C_5 and C_6. As has already been clarified during the study of the operation of the previous type of discriminator, the presence of frequency modulation in the signal causes amplitude changes of the high-frequency voltages fed to diodes D_1 and D_2. Hence, the voltages appearing across capacitors C_3 and C_4 also change at the frequency of modulation, but the algebraic sum of these voltages remains constant. What happens is that a redistribution of voltage takes place between these capacitors during modulation. It is evident that during this process the output voltage also varies at audio frequency.

As follows from the above, if the signal has a certain amount of amplitude modulation, such modulation will not affect the output voltage of the discriminator and, therefore, there is no need to provide a special detector preceding a discriminator of such type. Filter R_3C_7 accounts for a certain attenuation of the higher audio frequencies. These frequencies are, however, usually emphasised at the radio broadcasting station anyway, and, therefore, there is no loss of fidelity in this type of discriminator.

Fig. 250 gives the circuit diagram of the so-called "phase discriminator", which may be employed for the detection of *FM* signals both in superheterodyne and straight-amplification receivers. In this type of discriminator, the input-

signal voltage is fed from tuned circuit L_1C_1 to the third grid of a heptode valve, into the grid *1* circuit of which is connected tuned circuit L_2C_2, adjusted to the carrying frequency (to the average frequency) of the signal. Proper operation of this discriminator calls for high Q of tuned circuit L_2C_2. This is easier to obtain if the discriminator is fed with intermediate-frequency oscillations, which have a lower frequency than the frequency of the incoming signal. Tuned circuits L_1C_1 and L_2C_2 must not have any external coupling to each other. They are coupled only by the common electron stream within the valve.

The signal voltage, reaching the third grid of the valve, causes pulsations of the electron stream. This stream pulsates at the frequency of the signal and gives rise to a certain alternating e.m.f. in tuned circuit L_2C_2. When the signal is at its average value, the voltage across tuned circuit L_2C_2 lags by 90° behind the voltage across tuned circuit L_1C_1, thus lagging by 90° behind the voltage at the third grid. But then the negative voltage at the third grid and at the first grid will act, in general, only during three-fourths of each period (first-grid voltage acting during one quarter of each period, first-grid and third-grid voltages acting during the second quarter of the period, and third-grid voltage acting during the third quarter of the period). In this case, the anode-current pulses will have a certain average duration. If the frequency of the signal increases, then the phase shift will also increase and the duration of anode-current pulses will decrease, because the negative voltage at the grids acts longer than three-quarters of a period. For instance, when the phase shift is equal to 180°, the negative voltage acts during the entire period—acting during one half-period at one grid, and during the next half-period at the other grid.

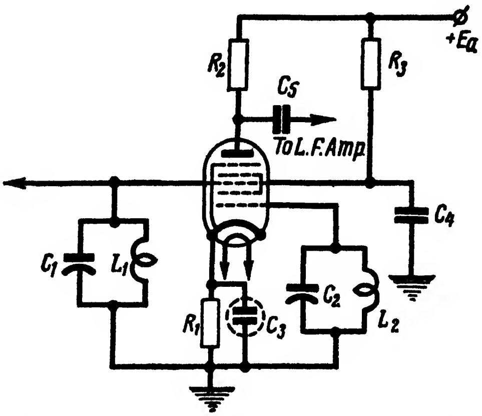


Fig. 250. The circuit diagram of the phase discriminator designed for the reception of FM signals

If the signal frequency decreases, the phase shift will be reduced.

In this case, the negative voltage at the grids will exist during a shorter span of time than three-quarters of a period (for instance, only during one-half of a period when the phase shift is zero). Consequently, in this case, the duration of anode current pulses will be increased. As a result, the average value of the anode current changes in accordance with the frequency deviation of the incoming signal. The resultant audio-frequency voltage appears across load resistor R_2 and is next fed to the input circuit of a l.f. amplifier.

The phase discriminator does not possess any amplitude-limiting properties and must be, therefore, preceded by a limiter stage.

118. AUTOMATIC FREQUENCY CONTROL.

The dependability of radio communication is enhanced by the incorporation of crystal control in the radio transmitter. It is difficult, however, to secure the frequency stabilisation in the transmitter over a continuously-variable band. During recent years the dependability of communication has been, therefore, secured over continuously-variable bands and at all frequencies through the use in the receiver of what is known as the automatic frequency control. The

principle of the automatic frequency control (generally abbreviated as a.f.c.) is shown by the block diagram in Fig. 251. (It should be kept in mind that, apart from the question of transmitter stability, the stability of the receiver heterodyne is also of paramount importance, as far as the dependability of communication is concerned.—*Translator's note.*)

A radio receiver provided with an a.f.c. circuit possesses, besides its usual stages, also a discriminator stage and a special controlling stage. The discriminator can employ any suitable circuit, for instance that shown in Fig. 248*a*. This discriminator stage is supplied with alternating voltage from the i.f. amplifier. When the frequency of this voltage is exactly equal to the resonant frequency of the tuned circuits of the discriminator, the d.c. output voltage at the output of the discriminator stage is equal to zero. When the signal voltage

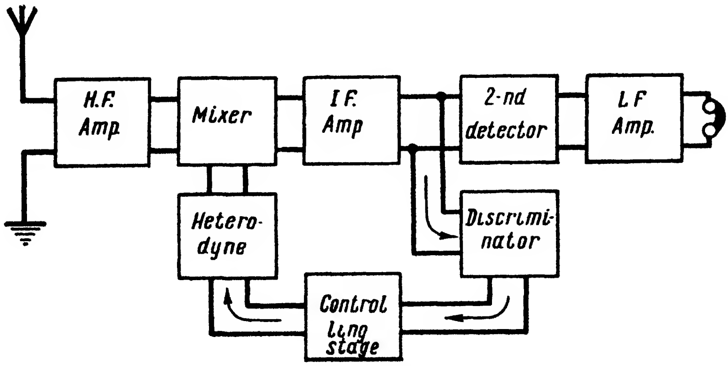


Fig. 251. The block diagram of a superheterodyne receiver employing an a.f.c. circuit

frequency deviates in either direction, an output voltage of respective polarity will be developed at the discriminator output. This d.c. voltage is fed to the controlling stage, this stage being similar to the reactance-valve modulator discussed under Sec.96. When the bias voltage at the grid of the reactance modulator valve varies, the inductance of the valve between the anode and cathode varies, too. The controlling valve is connected to the tuned circuit of the receiver heterodyne and controls the frequency generated by the heterodyne.

Let us assume that the frequency of the incoming signal has decreased as a result of certain instability of the transmitter. As a result, the frequency of the signal passing through the i.f. amplifier of the radio receiver will be increased. (The same effect will take place when the transmitter frequency is stable, but the frequency of the receiver heterodyne increases as a result of receiver instability.)

Analysing the above case, we see the following. The increase of the frequency of the voltage applied to the input circuit of the discriminator results in the appearance of a negative d.c. voltage at the output of the discriminator stage. This voltage, acting upon the grid of the controlling valve, will reduce the mutual conductance of the given valve. As a result, an increase of inductance will take place, the inductance being equivalent to the anode-cathode section of the controller valve. Since this inductance is connected in parallel with the tuned circuit of the heterodyne, the frequency of the voltage generated by the heterodyne will be decreased. This will decrease the difference between the heterodyne frequency and the frequency of the signal and the intermediate frequency will have a tendency to return to its normal value. If the frequency of the incoming signal has changed in the opposite direction (increased), resulting in a decrease of the intermediate frequency — which can also be decreased as a result of the

receiver heterodyne instability — the described process will be repeated. Only, in this case, a positive voltage will be developed at the output circuit of the discriminator. This will decrease the inductance value of the anode-cathode section of the controlling valve, resulting in an increase of the heterodyne frequency.

As follows from the above description, should any difference appear between the frequency of the incoming signal and the frequency to which the receiver is tuned, the a.f.c. circuit will act automatically to change the heterodyne frequency and maintain the stability of the intermediate frequency.

119. QUESTIONS AND PROBLEMS

1. Why does a radio receiver need high amplification?
2. What is the selectivity of a radio receiver?
3. Why does a crystal receiver provide a comparatively low level of audio signal?
4. The sensitivity of one radio receiver is 100 microvolts, while the sensitivity of another receiver is 5 microvolts. Which one of these two sets has a higher sensitivity?
5. Why should not the sensitivity of a radio receiver be made too high?
6. When a certain radio receiver is detuned by 10 kc from the frequency of an incoming signal, the level of the reproduced signal is decreased by 10 times. Does this receiver have a sufficiently high selectivity?
7. A radio receiver is capable of passing a frequency band of 300-2,000 cps. Is such a band sufficient for a satisfactory reproduction of music?
8. What kind of selectivity curve represents the satisfactory performance of a radio receiver?
9. Is the detection process in a radio receiver the same thing as the frequency conversion?
10. Are the detection and the rectification of alternating current similar processes in a radio receiver?
11. In a straight-amplification receiver, type 1-V-2, the first stage provides an amplification of 8, the second stage — 100, the third and the fourth stages —15 each. Determine the overall gain of the receiver.
12. Why must the load resistor of a diode detector have a high value?
13. Will the diode detector function properly if its load resistor is shunted by a 25,000-pf capacitor?
14. Explain why a diode detector will stop functioning if a steady positive d.c. potential is applied to the anode of the detector valve.
15. What is the similarity and the difference between the grid detector and the diode detector?
16. Will the grid detector function if a negative bias is applied to its control grid?
17. Which type of detector is most suitable for the reception of weak signals?
18. What is the difference between a self-excited valve oscillator and a regenerative detector?
19. The carrier frequency of a radio-telegraph transmitter is 500 kc. To what frequency must be adjusted the tuned circuit of the regenerative detector, picking up this carrier, so that the frequency of the reproduced audio note is equal to 1,500 cps?
20. Why does the feedback circuit of a radio receiver help to amplify high-frequency oscillations but not low-frequency oscillations?
21. Why is it desirable to provide a h.f. amplification stage in a regenerative receiver?
22. A regenerative stage of the receiver fails to oscillate, even when its feedback control is set to the maximum value. What could be the causes of such a failure?

23. What are the shortcomings of a h.f. amplifier stage employing, as a load, a choke or a resistor instead of a tuned circuit?

24. Why is it that triode valves are not used in h.f. amplifier stages?

25. Is it necessary to employ bias voltage in h.f. amplification stages?

26. Draw the circuit diagram of a regenerative type 1-V-0 straight-amplification receiver. Both stages are to use pentode valves. A tuned anode circuit is to be provided at the output of the h.f. amplifier. A variable capacitor is to control the regeneration.

27. A regenerative receiver is picking up a 300-kc c. w. signal. The tuned circuit of the regenerative detector is brought up to the threshold of oscillation and is tuned, in succession, to the following frequencies: 295 kc, 299.2 kc, 301 kc. What will be the frequencies of the audio beats in the earphones, of the receiver for each one of the three positions of the tuning control?

28. Devise the circuit diagram of a battery-powered, type 0-V-1, straight-amplification receiver, in accordance with the following design specifications. The first stage is to use a triode grid detector. The second stage (l.f. amplifier stage) is to employ a pentode valve directly coupled to its load. The interstage coupling is to be provided by means of transformers. The second stage of the receiver is to be self-biased.

29. Devise the circuit diagram of a radio receiver, specifying the values of all the components. The receiver is to be a type 1-V-1 straight-amplification set employing valves 6K7, 6C5 and 6П6C and is to incorporate the following features: 1) inductive aerial coupling; 2) parallel anode feed in the h.f. amplifier stage; 3) capacitor-controlled feedback in the detector stage (the second stage); 4) resistance coupling between the detector and the l.f. amplifier; 5) transformer output; 6) independent automatic biasing in the first and the third stages; 7) dropping resistors in the screen-grid circuits; 8) decoupling filters in the anode circuits of the first two stages.

30. Why is the sensitivity of a superheterodyne receiver higher than that of a straight-amplification set?

31. The intermediate frequency of a superheterodyne receiver is 470 kc, and the set is tuned to a 3,600-kc carrier. What must be the frequency of the receiver heterodyne to make the set reproduce the signal?

32. Does the intermediate frequency of a superheterodyne receiver vary when the ganged tuning capacitors of the set are rotated during the reception of a radio station?

33. Why is it that a superheterodyne receiver does not reproduce any signals when its heterodyne stage is not functioning?

34. What is the difference between a frequency converter and a mixer?

35. Why the broadcast radio receivers do not employ beat-frequency oscillators?

36. Why does a superheterodyne receiver possess a higher selectivity than a straight-amplification set?

37. The intermediate frequency of a superheterodyne receiver is 120 kc. The set is tuned to a 5,420-kc signal. What is the carrier frequency of the radio station creating image interference to the reception under the above-stipulated conditions?

38. The signal of a radio station operating at a frequency of 7,800 kc is reproduced by a radio receiver, when the heterodyne frequency of the set is equal to 8,260. At what other frequency of the heterodyne will the receiver reproduce the same signal?

39. What is a band filter?

40. What is the operating principle of the automatic gain control?

41. Can the automatic gain control circuit eliminate variations of the strength of a signal reproduced by a radio receiver, if such variations are caused by signal fading?

42. How do you understand the effect of interference reduction with the help of a manual gain control?

43. Is the "magic eye" an indispensable part of a radio receiver?

44. On what wavelength ranges is the atmospheric interference most pronounced?
45. What types of electrical equipment create the strongest interference to radio reception?
46. Why is the effect of receiver noise more pronounced in superheterodyne receivers than in straight-amplification sets?
47. How would you suppress the man-made electrical noise at the source of such noise?
48. Explain the operating principle of the "noise-suppressing" aerial.
49. Devise the circuit diagram of a superheterodyne receiver, employing valve 6A7 in the frequency converter stage, valve 6K7 in the i.f. amplifier stage, valve 6Г7 in the detector and the 1st l.f. amplifier stages, and valve 6П6С in the output stage. The receiver is to have a manual gain control and a tone control. All the stages of the receiver are to be automatically biased. The first two stages are to incorporate anode decoupling filters. Show the approximate specifications of all the components of the receiver circuit.
50. In what cases is it advantageous to decrease the selectivity of a radio receiver?
51. What is the principle of reception of frequency-modulated signals?
52. Find the gain of a high-frequency amplifier stage in which the valve parameters are as follows: $\mu = 2{,}000$ and $R_i = 2$ megohms, while the equivalent resistance (R_e) of the tuned anode circuit is equal to 20,000 ohms.
53. What are the shortcomings of a superheterodyne receiver, in which the heterodyne frequency is lower than the frequency of incoming signals?
54. What is the role played by beat frequencies in the radio reception technique?
55. Is it good practice to connect a gramophone pickup to the grid of the i.f. amplifier valve or to the diode detector of a superheterodyne radio receiver?

CHAPTER X

RADIO MEASUREMENTS

120. CURRENT MEASUREMENTS

Measurement of Direct Current

It often becomes necessary to measure direct current in supply circuits of electron valves, the current value characterising each particular operating condition of a valve.

Permanent-magnet moving-coil instruments, chiefly milliammeters, are employed for the above type of measurement. Let us first recall that the permanent-magnet moving-coil instruments (usually referred to simply as "moving-coil meters") are suitable for direct-current measurements only, and, accordingly, may be connected into a circuit only when the polarity of connection is correct. The only exception, in this case, are those moving-coil meters in which the zero position of the pointer is in the middle of the scale; the polarity of connection of such meters is immaterial, as long as the meter is connected into a d.c. circuit.

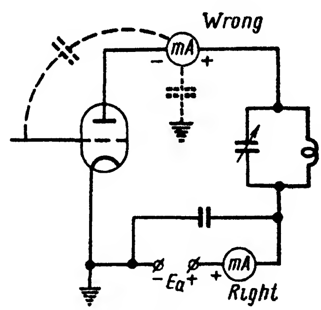


Fig. 252. Right and wrong ways of milliammeter connection in the measurement of the d.c. component of the anode current of a valve

Moving-coil meters are more sensitive than other types of measuring instruments and their scales are calibrated in equally-spaced divisions. The most sensitive varieties of moving-coil meters are called galvanometers and microammeters. These instruments can read extremely low currents (some of them can measure currents as small as a fraction of one-millionth part of one ampere). It is only natural that such meters require very careful handling, or else they can be easily damaged. Speaking generally, all types of moving-coil meters must be handled with care and protected against overloads. Because of this, before attempting to make a measurement, an approximate value of the current to be read by the meter must first be ascertained. This done, the meter is switched on to give a reading on its higher scale (i.e., on the higher measurement range). Only after this, the instrument may be switched over to an appropriate scale. If the meter reading is too low on the given scale, the instrument may be switched over to a lower scale (i.e., to a more sensitive range). In this way it is possible to use a single sensitive instrument for the measurement of currents of greatly varying values; it is only necessary, in this case, to provide a number of different shunts for the given meter, each shunt designed for a definite measurement range. A rotary switch, connecting a required shunt in parallel with the meter, can be

incorporated in such a meter, making of it a universal measuring instrument. The highest precision of measurement requires that the shunts are wound with some type of high-resistance wire (manganin, constantan, nickelin, nichrome, etc.).

When the direct current flowing in the anode circuit of an electron valve is to be measured, the meter should be so connected that no a.c. component flows through it. Fig. 252 illustrates a wrong and a correct way of connecting the meter in this case. If the meter is connected at the anode, parasitic capacitances are created (broken lines in the drawing), the values of these capacitances being determined by the size and dimensions of the instrument and by the length of connecting wires. This capacitance and the distributed inductance of the meter can upset the correct operating condition of the stage. This is particularly apt to happen at high frequencies and in some cases can result in parasitic oscillation of the stage. Should this occur, the value of the d.c. component of the anode current will be changed and the measurement will be erroneous.

Low alternating currents may be safely passed through a moving-coil meter. However, when such meters are installed in high-power or in high-frequency equipment (for instance, in transmitters), where large alternating currents flow in the d.c. circuits or in their vicinity, the meters must be protected from alternating currents. High-capacitance capacitors, connected in parallel with the meters, provide the necessary protection.

Measurement of Low-Frequency Alternating Current

50-cps current is usually measured by means of moving-iron instruments, although other types of meters (thermal, thermocouple, electrodynamic instruments) are occasionally used for the purpose. Uneven calibration of the scale is the common shortcoming of all alternating-current measuring instruments. The divisions are made smaller at the beginning of such a scale. This limits the measurement of low currents to such an extent that in practical a.c. meters the smallest current reading that can be taken is equal to about 20% of the maximum current value read at the end of the scale.

Moving-iron instruments are manufactured with scales ranging from one ampere (full-scale reading) and higher. These meters cannot measure high-frequency currents because the meters possess a large inductance and a considerable distributed capacitance. However, the given instruments may be employed for the measurement of direct current, although their scales are denoted with a special mark (~) standing for alternating current only.

Thermal and thermocouple meters may be employed for the measurement of low-frequency current of any frequency. These instruments can be designed for higher sensitivity than the moving-iron meters. Formerly, the now absolute thermal meters used to be manufactured with scales ranging up to 100 milliamperes. These meters have been completely replaced by the thermocouple meters, whose sensitivity is even higher than that of the thermal instruments.

Low-frequency measurements can also be made by means of the so-called rectifier-type instruments. In these instruments the alternating current is rectified into a pulsating current by semiconductor rectifiers, and then the d.c. component of the pulsating current is read on the scale of the usual moving-coil instrument, normally designed for d.c. service only.

The circuit of a rectifier-type measuring instrument should be normally so arranged that both half-waves of the rectified current flow through the meter. If the meter is designed in this manner, its connection into a circuit whose current is to be measured will not upset the operation of such a circuit. The only exception from this rule (when a single rectifier unit may be used) is permitted in the case when nearly the whole of the rectified current is passed through the shunt (Fig. 253*a*). In the circuit shown in Fig. 253*b* a single rectifier unit R_1 is connected in series with the measuring instrument (providing a half-wave

rectification), but the opposite half-waves of the current are made to flow through another rectifier unit R_2. In this circuit, R_2 is connected in series with resistor R, whose resistance is equal to resistance R_m of the meter. The sensitivity of an alternating-current measuring instrument may be increased by the application of the usual transformer-type full-wave rectification (Fig. 253*c*), or else by a bridge circuit with four rectifier units (Fig. 253*d*). The latter system is more frequently used.

High sensitivity (i.e., the possibility of measuring very small currents) is the advantage offered by the rectifier-type measuring instruments. When these instruments employ half-wave rectification their sensitivity on alternating-current measurements is 2.5-3 times smaller than the sensitivity of moving-coil instruments on direct current measurements. With full-wave rectification, the

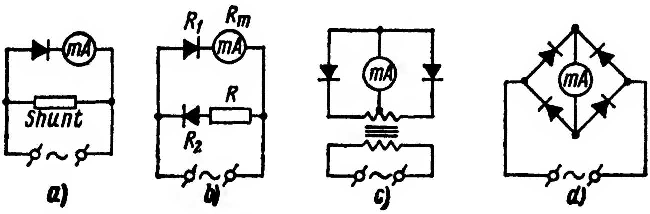


Fig. 253. Circuit diagrams of rectifier-type current-measuring instruments

sensitivity of the a.c. measuring instruments is only 25-50% smaller than that of the d.c. instruments. For instance, if a 1-ma moving-coil meter is employed by the circuit of Fig. 253*b*, the meter will give a full-scale deflection when the whole measuring instrument consumes 2.5-3 ma. The incorporation of the same meter in the circuit of Fig. 253*d* will provide the full scale deflection when the consumption of the current is equal to 1.25-1.5 ma. The measurement range of the a.c. instruments may be broadened by connecting appropriate shunts across the terminals to which the alternating current under measurement is applied.

The rectifier-type measuring instruments just described are noted for several shortcomings. First, they possess a considerable resistance, and, because of this, account for a certain voltage drop (at least some fractions of one volt) when the current being measured flows through them. Secondly, although the scale divisions of these instruments are fairly equally spaced, still there is some compression of these divisions at the beginning of the scale. This peculiarity is quite confusing in multi-range meters of the given type, because the non-equality of division-spacing varies on different measurement ranges. Thirdly, the rectifying properties of the rectifier units used in these instruments depend upon the temperature and frequency; consequently, the accuracy of the instrument varies at different ambient temperatures and on different frequencies of the current being measured. Another disadvantage of these instruments is the considerable capacitance of their rectifier units, this capacitance easily by-passing currents of the higher frequencies. This also affects the accuracy of the instruments. In case of cuprous-oxide rectifier units, such capacitance reaches 0.03 mfd to each sq cm of the contacting surface. Because of this peculiarity, the usual high-capacitance cuprous-oxide or selenium rectifiers may be used in the described type of instruments only on frequencies below 50 cps. The measurement of higher-frequency currents requires the incorporation of special low-capacitance cuprous-oxide rectifier units or germanium diodes in the measuring instruments.

The readings of the rectifier-type measuring instruments depend upon the shape of the current being measured. These meters are usually calibrated on

sinusoidal current, and, therefore, give erroneous readings when the currents of other shapes are measured. Concluding the list of shortcomings of these instruments, it should be noted that the rectifying properties of some cuprous-oxide rectifiers change with the passage of time, thus varying the accuracy of meter readings.

Measurement of High-Frequency Current

High-frequency current is measured with the help of the thermal and thermocouple instruments. These instruments offer good operation on frequencies up to several megacycles. After this frequency limit is reached, the skin-effect and the effect of the parasitic capacitances begin to tell. As already noted above, the thermocouple instruments have at present almost completely replaced the instruments of the thermal type.

Fig. 254 gives the basic circuit diagrams of the thermocouple instruments. In the circuit diagram shown in Fig. 254*a*, the high-frequency current being measured flows through wire *H* (called the heater) and heats the junction of two dissimilar conductors *TC*. These conductors are called the thermocouple and are connected to a moving-coil meter (usually a galvanometer or a milliammeter). The thermocouple together with the heater constitutes what is known as the thermoelement. If the thermocouple contacts the heater, it is known as a *contacting thermocouple*. The advantage of a contacting thermocouple is its high-sensitivity, its disadvantage — the branching off of the high-frequency current through the contact to the d.c. circuit. The latter circuit usually has considerable parasitic capacitance to the earth (i.e., to the chassis of the instrument), as shown by the broken line in Fig. 254*a*. The circuit shown in Fig. 254*b*

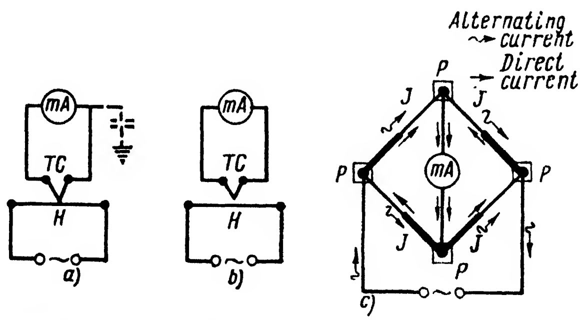


Fig. 254. Circuit diagrams of thermocouple instruments

and employing a *non-contacting thermocouple* is almost completely free of this defect, because in the given circuit the capacitance between the heater and the thermocouple is quite small. On the other hand, in the non-contacting thermocouple a smaller heat transfer takes place between the two elements because of the insulation placed between them. As a result, the sensitivity of this type of thermocouple is lower than that of the contacting thermocouple.

In order to increase the value of direct current flowing through the galvanometer, it becomes necessary to obtain a higher value of e.m.f. from the thermocouple and to lower the resistance of the thermocouple and of the galvanometer. However, low-resistance galvanometers are usually not very sensitive and, because of this, it is not advantageous to employ such meters. As far as the e.m.f. is concerned, it can be increased by connecting several thermocouples in series, but this also brings up its own difficulties, because such a connection increases

the resistance of the thermocouple circuit. The latter shortcoming is eliminated in the bridge-type thermoconvertor (Fig. 254c), where four thermocouples are arranged in a bridge circuit. The high-frequency current is applied across one diagonal of the bridge and flows along two branches, heating the thermocouples. These thermocouples, acting as the sources of e.m.f., are arranged in two parallel groups, each group consisting of two series-connected thermocouples. The effect of such connection is the doubled value of e.m.f., the internal resistance of the whole combination remaining at the same value as the internal resistance of a single thermocouple. The paths of direct current and of the high-frequency current are shown in Fig. 254c. In the bridge circuit, only junctions J should be subjected to heating. All the other junctions must remain cool, the cooling being provided by metal plates P. In order to reduce the heat loss, the thermoconvertor is sometimes placed in a small evacuated glass envelope. The thermocouples are usually comprised of iron-constantan pairs. Such a pair may be heated to a temperature as high as 600°, developing an e.m.f. of 30 mv at such a temperature.

In comparison with the thermal instruments, thermocouple instruments provide a better sensitivity. Besides, their reading error is smaller on high frequencies. The latter is attributed to the fact that the heater of the thermocouple instrument is shorter than the hot-wire of the thermal instrument and, hence, possesses smaller inductance and capacitance. Another advantage of the thermocouple instrument is that it is more convenient in use, because the galvanometer can be installed even at a large distance from the thermocouple unit by lengthening the wires carrying only the direct current. It should be noted, however, that this type of instrument must be handled with care; the thermocouple will be easily damaged if excessive current is passed through it.

Another thing to bear in mind is that on high frequencies the measuring instruments should not employ shunts. On these frequencies, the impedance of the shunt is slightly different for each particular frequency due to the varying values of distributed capacitance, inductance and skin-effect in the shunt.

In radio engineering practice, simple high-frequency indicators are often used when it is required to obtain only an approximate current measurement. The usual low-voltage miniature incandescent lamp is a typical indicator of such kind. The best measurement accuracy with this type of indicator is obtained when the lamp just begins to glow. The lamp must first be calibrated on direct current in order to establish the current value at which the glow begins. A typical reference value, in this case, may be given by a 3.5-volt torchlight lamp; this lamp, although designed for normal operation on 0.25 a, begins to glow when a current of 0.1 a flows through it. When it is required to measure a greater current than 0.1 a, it is only necessary to take several such lamps and connect them to a rotary switch in such a way that different positions of the switch rotor will connect in parallel as many of the lamps as may be necessary. The switch is set to such a position where the lamps connected by it to the circuit just begin to glow. This gives a fairly good idea of the current flowing through the combination; for instance, if the switch has connected into the circuit three lamps, which had just began to glow, the total current flowing through the lamps is equal to about 0.3 a. This method should be used with a certain caution on high frequencies, because the lamp bases possess considerable capacitance. As a result of this, a part of the high-frequency current will flow not only through the lamp filament but through the said capacitance. This will lead to considerable reading errors, which can be reduced only by removing the lamp bases before the lamps are used for high-frequency measurements in such a circuit.

Miniature incandescent lamps are frequently employed to indicate resonance by connecting them in series with tuned circuits. In this case, the lamps are not intended to measure the current but simply to show that the circuit is in its resonant condition. Such condition is indicated by the maximum brightness of the glow, corresponding to the greatest value of current flowing through the lamp. If the current flowing through the tuned circuit is great enough to cause a burn-out of the lamp, the lamp may be shunted with a piece of wire,

or else, inductively coupled to the tuned circuit. When the indicator of this type is to be connected to the aerial circuit of a radio transmitter (the current in this circuit is usually considerably greater than the maximum current that can be passed safely by the lamp), the lamp should be connected in the following manner to indicate the resonant condition of the aerial. First, a small loop is made in the aerial wire. Then, a short piece of independent wire is coiled to form several turns and this coil is inductively coupled to the loop of the aerial wire. Lastly, the two ends of the coil are soldered to the lamp terminals — and the indicator is ready for use.

When the transmitter is switched on, the lamp will glow. A word of caution is in order in this case, however. The coil should be initially coupled very loosely to the loop in the aerial. If the coupling is close, and the transmitter power is high, switching on the transmitter is likely to blow the lamp. The proper way to proceed is as follows. Beginning with a very loose coupling between the aerial

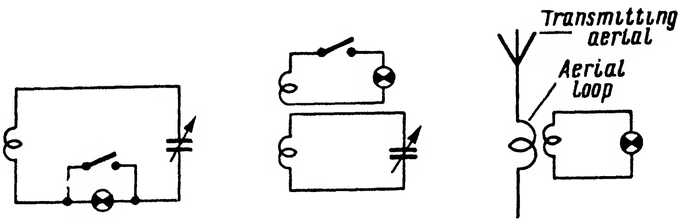


Fig. 255. Circuits of current indicators employing incandescent lamps

loop and the coil, the transmitter is switched on and the coupling is gradually made closer until the lamp just begins to glow when the aerial is adjusted to resonance. The coupling then may be fixed in this position without any danger of blowing the lamp, because the aerial current value will never be greater than the value obtained at resonance.

When the described procedure is followed, the lamp indicator may be safely employed with radio transmitters of any power.

It should be noted here that the lamp indicator draws a certain amount of high-frequency power from the aerial. Because of this, and particularly when dealing with low-power transmitters in which the aerial power is at a premium, the lamp must be moved away from the aerial, once the aerial circuit has been tuned to resonance. In some low-power transmitters the lamp is connected right into the aerial wire for the purpose of tuning. In such transmitting sets the lamp should be short-circuited, once the condition of aerial resonance has been obtained. Typical lamp indicator circuits are given in Fig. 255.

Calibration of Current-Measuring Instruments

Before any type of current-measuring instrument is put into practical application, its scale must be correctly calibrated, otherwise the readings of the instrument may be erroneous and, hence, quite worthless.

There are two general ways of calibrating current-measuring instruments, as illustrated in Fig. 256.

In the circuit shown in Fig. 256*a*, a standard instrument A_s (ammeter or milliammeter) is used to measure the current. This standard meter is connected in series with instrument A, which is to be calibrated. An additional resistor, limiting the current-flow through the two series-connected meters, is also inserted into the circuit. When the calibration of meter A is correct, the readings of the two meters will coincide.

In the circuit shown in Fig. 256*b*, a standard voltmeter is employed, while resistor R possessing a known resistance value is connected in series with the instrument A (ammeter or milliammeter under calibration). This value is much greater than the resistance values of meter A. In this circuit, the correct value of current is determined by dividing voltage U by the value of R. (If the value of R is comparatively not high, it is best to connect the standard voltmeter in parallel with this resistor.)

In either one of the above circuits, the calibration process is carried out as follows. As can be seen from the diagram, the calibration circuit is supplied with power from a potentiometer. The value of current flowing through the instrument under calibration is varied by moving the potentiometer-slider. The slider is moved to a certain position. In this position, the standard instrument is referred to in order to determine the value of current flowing through the meter under calibration. The reading of this meter is noted and the reading of the

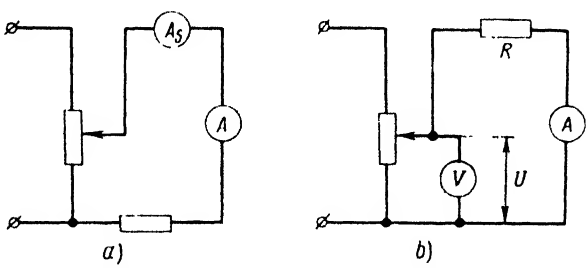


Fig. 256. Calibration circuits for current-measuring instruments

standard instrument is recorded. This operation completed, the potentiometer pointer is moved to another position, where a different value of current flowing through the meter under test is established. This done, the value of current flowing through this meter is again noted and the respective reading of the standard instrument is recorded. The described process is continued for a number of settings of the potentiometer-slider, and all the readings obtained are entered into a table. The scale of the meter under test is then calibrated in accordance with such table.

When the meter being calibrated is intended to operate on alternating current, it is a good practice to carry out the calibration on those frequencies on which the current values will be measured by the meter. This, however, is not always possible to predict. Hence, the calibration of such a meter is generally carried out on direct current or else on the industrial 50-cps current.

Rectifier-type instruments, as well as bridge-type thermocouple instruments (Fig. 254*c*) must be calibrated on alternating current. If no a.c. calibration facilities are available, a thermocouple instrument provided with a contacting thermocouple may be calibrated on d.c. if the following procedure is used. For every value of current flowing through the heater, two readings of the meter under calibration are taken—one, for the current in one direction, and the other, for the same value of current in the reverse direction. When this is done and the two readings of the meter for any particular value of current are not equal, the average of these readings is taken to represent the current value.

In the above discussion, the calibration circuit was supplied with power from a potentiometer. Alternative methods of power supply are also possible. Thus, when a.c. instruments are being calibrated, the calibration circuit may be fed with power from the mains through a step-down transformer. A rheostat connected in series with the primary winding of such a transformer will provide the necessary voltage adjustment.

121. VOLTAGE MEASUREMENTS

Measurement of D.C. Voltages

In radio engineering practice it often becomes necessary to measure the voltage of power supply sources, as well as various d.c. voltages at valve electrodes. The operating condition of every electron valve is determined by appropriate values of voltage applied to the valve anode, control grid, screen grid, etc. Hence, the measurement of these voltages is very important for the adjustment of various radio equipment, for checking the operating condition of individual stages, and for location of the causes of faults in the equipment. When it is required to measure d.c. voltages, moving-coil voltmeters are generally used. However, the usual types of voltmeters of this kind, such as those used in power engineering practice, are not suitable for all types of d.c. radio measurements.

A voltmeter should be always connected between those points of an electric circuit between which the voltage is to be measured. The device must consume only very low current, i.e., its internal resistance must be as high as possible. The current consumption and the resistance are the main indices determining the suitability of the instrument for various types of radio measurements. Common types of voltmeters draw several milliamperes of current to obtain a full-scale deflection. This means that a device of this type has a resistance of a few hundred ohms to each scale division. For instance, when the voltmeter possessing a 0-1 v scale gives a full-scale deflection on a current-flow of 5 ma through the instrument, the resistance of such a voltmeter is equal to 200 ohms per volt. If this meter is redesigned to operate with a 3-volt scale, its resistance will be 600 ohms. For operation with a 100-volt scale, its resistance will be 20,000 ohms, etc. Voltmeters of the above type are known as low-resistance voltmeters.

Any milliammeter may be converted into a voltmeter. To do this, it is merely necessary to connect a series resistor of appropriate value to the milliammeter. Since the internal resistance of a milliammeter is negligibly small, the value of the series resistor is determined by Ohm's law in accordance with the voltage for which the voltmeter must be designed and in accordance with the current consumed by the milliammeter. For example, take a milliammeter designed to give a full-scale deflection when a current of 6 milliamperes flows through the meter coil. If such a milliammeter is to be converted into a voltmeter with a 300-volts scale, it is only necessary to divide 300 v by 6 ma. The result—50,000—will be the resistance value in ohms which must be possessed by the series resistor. When it is required to convert a single milliammeter into a multi-range voltmeter (i.e., into such a voltmeter which will give full-scale deflection on several ranges, for example, on 3 volts, 30 volts, 3,000 volts, etc), it is necessary to calculate appropriate series resistors for each indicated full-scale deflection. Connecting the corresponding resistor in series with the milliammeter for any particular voltage range gives a voltmeter capable of measuring low, medium and high voltages.

Low-resistance voltmeters may be used for voltage measurements only between such points of an electric circuit which are directly connected to the terminals of a power supply, and when there are no additional high resistances of any kind between the instrument and the power supply. Moreover, in this case the power supply itself must possess low internal resistance. For instance, a voltmeter of this type may be connected directly to the terminals of a storage battery and will give a correct reading, because the internal resistance of the battery is very small. In those electron-valve circuits, in which the valve electrodes are connected to their appropriate power supplies through low-resistance circuits, a low-resistance voltmeter may also be employed for fairly correct voltage measurements. Fig. 257*a* gives some examples where a low-resistance voltmeter may be employed for voltage measurements in an electron-valve stage. Apparently, an instrument of this type may be used to measure the anode voltage of a transformer-coupled amplifier, of a high-

frequency amplifier or of an oscillator. The meter may be also used for the measurement of screen-grid voltage — if the screen-grid is directly connected to its power supply without any intermediate resistors. A measurement of control-grid voltage with this instrument is also possible if the grid voltage is derived from an independent power supply. In all of these listed cases, the additional current drawn by the meter will not cause any noticeable redistribution of voltages in the circuits under reference.

Fig. 257*b* pertains to the connection of a voltmeter to the screen grid and the anode in a valve stage in which high resistances are connected in series with these electrodes. In this case, a low-resistance voltmeter would give totally erroneous, greatly lowered, readings. The reason for this will be clarified by the following example. Assume that in the given stage, resistance R_a connected in

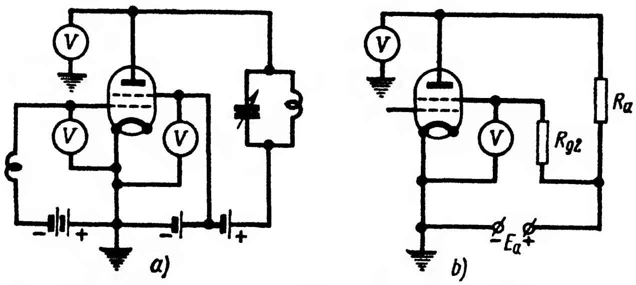


Fig. 257. Voltmeter connections for the measurement of d.c. voltages on valve electrodes

series with the anode circuit is equal to 100,000 ohms. Also assume that the operating condition of the stage is such that the anode current $I_a = 0.5$ ma, while voltage E_a of the anode power supply is equal to 200 v. As follows from the above conditions, anode voltage U_a of the valve will be equal to: $U_a = E_a - I_aR_a = 200 - 0.5 \times 10^{-3} \times 100{,}000 = 200 - 50 = 150$ volts.

In the case under consideration, the whole of the voltage developed by the anode power supply is distributed between the resistor R_a and the internal resistance R_0 offered by the valve to direct current, the value of R_0 being equal to $\frac{150}{0.5 \times 10^{-3}} = 300{,}000$ ohms. Apparently, the voltage drop across R_a is equal to 50 v, and across the valve it is equal to 150 v.

Now assume that a 150-volt low-resistance voltmeter, having an internal resistance $R = 30{,}000$ ohms (i.e., 200 ohms per volt), is connected between the anode and cathode of the valve. This will have the following effect. The voltage will now be distributed between resistances R and R_a, which are equal, respectively, to 30,000 and 100,000 ohms. (We can neglect the influence of R_0, connected in parallel with the voltmeter, because R_0 is much greater than R.) As a result, a voltage of 46 v will be set up across the voltmeter terminals, and, hence, also across the valve. Thus, a low-resistance voltmeter connected in the described manner will read 46 v as a value of the anode voltage, while, in reality, this voltage was 150 v before the voltmeter was connected with the circuit.

The above example makes it clear why low-resistance voltmeters cannot be employed for measuring the voltages at valve electrodes in resistance-coupled amplifiers, in screen-grid circuits with high dropping resistors, in circuits with decoupling filters, etc.

High-resistance voltmeters must be used for all measurements in such circuits, and these voltmeters must possess an internal resistance of at least 10,000 ohms per volt, i.e., they must not consume over 0.1 ma of current to give a full-scale deflection. Using a voltmeter of this type in the numerical example given above, it becomes apparent that the internal resistance of the instrument will

be 1,500,000 ohms on the required 150-volt scale. When such an instrument is connected in parallel with the valve, the connection will cause only an insignificant voltage redistribution and the meter will give a reading of only slightly less than 150 v.

A high-resistance voltmeter normally consists of a galvanometer (or mircoammeter) and of series resistors. Fig. 258 shows the circuit of such a voltmeter designed for multi-range operation (4 ranges). In this type of instrument, the series resistors may be of non-wire-wound type. If the galvanometer draws 100 microamperes and the voltmeter is provided with 5, 20, 100 and 500-volt scales, the series resistors will have respective values of 50,000; 200,000; 1,000,000; and 5,000,000 ohms.

It should be noted, however, that even a voltmeter of this type is not suitable for measuring the bias voltage right at the grid of the valve in a stage employing a high value of grid resistor, or a resistor of grid-decoupling filter (Fig.

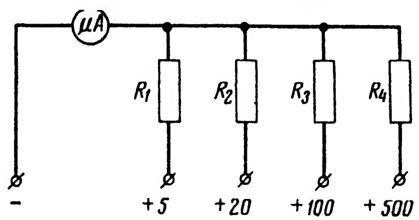


Fig. 258. The circuit of a high-resistance voltmeter employing a sensitive galvanometer

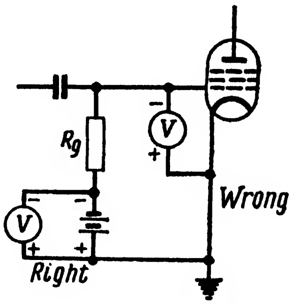


Fig. 259. Right and wrong ways of voltmeter connections when measuring the grid-bias voltage of a valve

259). Let us analyse the circuit of such a stage from the point of voltage measurements. Assume that a voltage of 5 v is fed through resistor R_g, the value of which is equal to 0.5 megohm. Let us now take a high-resistance voltmeter with a 5-volt scale and connect the instrument between the grid and cathode of the valve, assuming that the internal resistance of the instrument is equal to 50,000 ohms. Referring to the circuit diagram of Fig. 259, it may be seen that the meter would, in this case, be connected in series with the resistor R_g. This will cause a voltage redistribution between R_g and the internal resistance of the voltmeter, and the latter will give an erroneous reading of less than 0.5 v. Because of this, the voltmeter has to be connected not between the grid and cathode but rather right across the grid bias battery (as shown in the circuit diagram). If the valve is automatically biased, the voltmeter must be connected in parallel with the self-biasing resistor, i.e., between the cathode and the chassis.

A direct measurement of bias voltage at the control grid of a valve is possible only with the help of a valve voltmeter. Instruments of this type are discussed later.

Measurement of Low-Frequency Voltage

Moving-iron voltmeters are widely used for the measurement of low-frequency voltages in the circuits where the frequency does not exceed a few dozens of cycles per second. The moving-iron voltmeter is the usual type of instrument employed for a.c. power mains voltage measurements. This type of instrument is, however, unsuitable for voltage measurements in the range of audio frequencies because the meter coil is wound with a large number of turns and possesses a large distributed capacitance. Because of this, at audio-frequencies, the impe-

dance of the instrument widely changes with frequency. Besides, the moving-iron meter consumes a considerable current (at least 20 ma) and the resistance of the meter is not sufficiently high — (50 ohms per volt and even less) — to provide good reading accuracy.

Due to these shortcomings, moving-iron meters are not favoured in radio measurements, preference being given to rectifier-type voltmeters. The circuit diagrams of these voltmeters are similar to the circuit diagrams of rectifier-type current measuring instruments (Fig. 253), the only difference between the two kinds of meters is the inclusion of a dropping (series) resistor R_s in the a.c. side of the rectifier-type voltmeter (Fig. 260).

The simplest circuit diagram of a half-wave rectifier-type voltmeter employs only one rectifier unit (Fig. 260*a*). This circuit has an inherent disadvantage seen in that it does not pass the negative half-waves of the voltage being measured. This means that the full value of such voltage is applied across the rectifier unit during the negative half-waves, which can easily cause a breakdown of

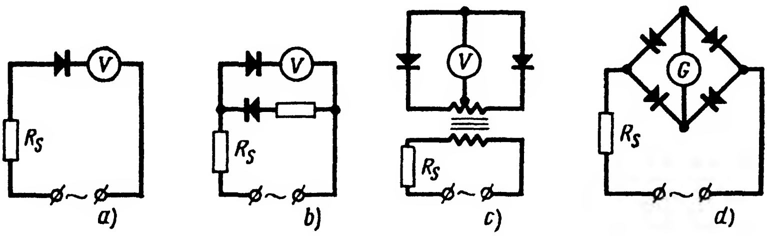


Fig. 260. Circuits of rectifier-type voltmeters

the rectifier unit if the voltage be high. Hence, the given circuit may be used only when the voltage being measured is lower than the working voltage of the rectifier unit. This limitation may be overcome by connecting several rectifier units in series in such an instrument; this would allow us to measure higher voltages. It is better, however, to use the circuit shown in Fig. 260*b*, where both the positive and the negative voltage half-waves are passed through the instrument. The circuit given in Fig. 260*a* also has another shortcoming. It is inconvenient in use, because the d.c. component of the current being measured must be passed through the circuit whose voltage is being measured. This is not always feasible, nor desirable.

Circuits shown in Fig. 260 *b*, *c* and *d* do not require a series connection of several rectifier units for the measurement of higher voltages, because in such circuits nearly the whole of the voltage being measured is dropped across series resistor R_s both during the positive and the negative half-waves. Among these circuits, the bridge-rectifier circuit, comprising four rectifier units, is considered to be the best.

The input impedance of a rectifier-type voltmeter can be increased only if a high-sensitivity galvanometer is employed. The value of such input impedance can be made as high as 6,000 or 8,000 ohms per volt in a bridge-rectifier voltmeter employing a 100-μa galvanometer. It is highly desirable that the natural capacitance of the rectifier units is made as small as possible when they are used in a voltmeter of this type. The value of series resistor R_s is calculated in a manner similar to dropping-resistor calculations in d.c. circuits. The result, however, should be divided by 2.5 or 3 in the case of a half-wave voltmeter and by 1.25 or 1.5 in the case of a full-wave instrument. Such calculation is, of course, only approximate, and gives a somewhat inflated value of R_s. Besides, the rectifier units themselves can have different resistance values, varying from some hundreds to many thousands of ohms. Because of these considerations, the exact value of R_s is found by the cut-and-try method during the calibration of the instrument.

A numerical example:

Calculate the value of R_s for a 150-volt scale to be employed by a bridge-type rectifier voltmeter employing a 200-μa galvanometer.

First, we find the value of R_s for the case when the circuit is used for d.c. voltage measurements:

$$R = \frac{150}{0.2\times10^{-3}} = 750{,}000 \text{ ohms.}$$

Now, dividing this resistance by 1.5, we obtain the final value of R_s, as follows:

$$R_s = \frac{750{,}000}{1.5} = 500{,}000 \text{ ohms} = 0.5 \text{ megohms.}$$

Apart from the rectifier-type voltmeters just discussed, low-frequency voltages can also be measured with valve voltmeters discussed in the following section.

Measurement of High-Frequency Voltage

Given below is the description of valve voltmeters, which are used for all high-frequency voltage measurements. These instruments are quite universal and may be also used for the measurement of d.c. voltages, as well as of low-frequency voltages, as stated above.

Going up into the radio-frequency, the valve voltmeter measurement range can be extended to measure voltages at frequencies as high as many dozens and even hundreds of megacycles. This is attributed to the fact that the instruments of this type possess a very low input capacitance C_i while their input resistance R_i can be made as high as several megohms, and even several dozens of megohms. These parameters of the valve voltmeter remain constant at all points of the frequency range of the instrument.

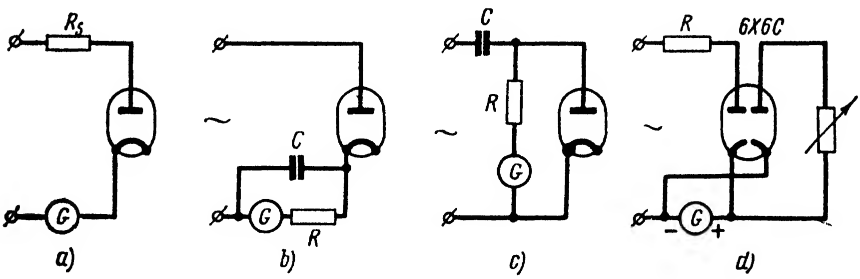


Fig. 261. Circuits of diode-valve voltmeters

Valve voltmeters of the diode type are the simplest. Diode-type valve voltmeters require no anode power supply, and this is their big advantage. However, the input resistance of these valve voltmeters is about the same as that of the rectifier-type voltmeters described above. When a diode-type valve voltmeter is used in conjunction with a microammeter, the input resistance of the instrument does not exceed some thousand ohms per volt.

The diode-type valve voltmeters are usually designed for half-wave rectification. The simplest circuit arrangement, in this case, is shown in Fig. 261*a*. The circuit is similar to the circuit given in Fig. 260*a*. The valve voltmeter, however, has an important advantage over the rectifier-type voltmeter in that it can withstand high negative half-wave voltages, and, besides, possesses a very low capacitance (a few picofarads). Series resistor R in the valve voltmeter

circuit is calculated in the same way as in the case of the rectifier-type voltmeter employing a half-wave circuit.

Valve voltmeters shown in Fig. 261 *b* and *c* are known as *peak voltmeters*. The circuits of these instruments are similar to the series and parallel circuits of diode detectors given in Fig. 208. The value of resistor R must be considerably greater than the value of anode resistance of the diode. If this is observed and capacitance C is sufficiently large, the d.c. voltage built up across capacitance C is almost equal to the amplitude value of a.c. voltage being measured, i.e., $U_{-} \approx U_m$ (or equal to the peak value if the shape of voltage is non-sinusoidal). The direct current flowing through the galvanometer is given by the following:

$$I_{-} \approx \frac{U_m}{R}.$$

Hence the value of resistor R should be computed from the following formula:

$$R \approx \frac{U_m}{I_{-}} = \frac{1.4U}{I_{-}}$$

where: U is the effective value of the a.c. voltage being measured, while I_{-} is the current corresponding to the full-scale deflection of the galvanometer. Thus, in the case of peak voltmeter circuits, R is made 40% higher, as compared to the case when a galvanometer is used for the measurement of d.c. voltages. The input resistance of the peak voltmeter circuit of Fig. 261*b*, is given as $R_i \approx \approx 0.5R$, while in the case of circuit of Fig. 261*c*, $R_i \approx 0.33R$. When the frequency is increased up to several megacycles and higher, the value of R_i is decreased, because various losses of energy and the shunting effect of the input capacitance begin to tell.

The circuit given in Fig. 261*b* is not protected from d.c. voltage, because it has an *open input circuit*, while the circuit with *closed input circuit* (Fig. 261*c*) may be used for the measurement of alternating components of pulsating voltages.

Capacitance C must be such that the capacitor discharges only slightly during one period of the lowest-frequency voltage. Besides this, the capacitance must exceed the interelectrode anode-cathode capacitance of the diode by at least 100 times. The following formula helps to calculate the least permissible capacitance value of C in picofarads:

$$C = \frac{20 \times 10^9}{R_{f\,min}}$$

where: f_{min} is the lowest frequency in cps, while R is given in thousands of ohms.

Numerical example:

Given a voltmeter, employing a circuit of Fig. 261*b*. The device is used in conjunction with a 100-μa galvanometer and is to measure effective voltage up to 20 v in the frequency range beginning with 100,000 cps. It is required to calculate the components of the given circuit. Proceeding as follows, we find:

$$R = \frac{1.4 \times 20}{0.1} = 280{,}000 \text{ ohms}; \quad R_i = 140{,}000 \text{ ohms};$$

$$C = \frac{20 \times 10^{-9}}{280 \times 10^5} = 700 \text{ picofarads}.$$

This is the least permissible value of capacitance, but the latter may be made larger than the calculated value. As may be seen, the voltmeter has a resistance of 7,000 ohms per volt.

The peak diode-type valve voltmeter is usually calibrated in effective values of sinusoidal alternating voltage. Then the amplitude value of sinusoidal voltage

may be found by multiplying the readings of such a voltmeter by 1.4. When the voltmeter is used to measure the effective value of non-sinusoidal shape voltages, the device will give erroneous readings because, in this case, the effective value is not equal to 0.7 of the peak value.

The diode-type valve voltmeters possess an insufficiently high value of input resistance R_i and, therefore, must be used in conjuction with very sensitive galvanometers. Besides, when this type of voltmeter is used to measure small values of low-frequency voltages, when R is comparatively small, it becomes necessary to employ a capacitor C of very high capacitance.

The presence of initial anode current I_{a0} in any circuit arrangement of the diode-type valve voltmeters is the disadvantage of these instruments. It should be noted that the initial current flows even when no a.c. voltage is applied to the anode of the valve. This initial current flow is chiefly attributed to the fact that some electrons are emitted by the cathode at a very high velocity, thus managing to reach the anode. This causes an initial deflection of the galvanometer pointer and, as a result, only a part of the galvanometer scale may be calibrated for a.c. readings. The effect of the initial current may be compensated for by connecting another diode into the circuit, making the current of this auxiliary diode pass through the galvanometer in the opposite direction. No alternating voltage is applied to this auxiliary diode, the initial current of which is adjusted by a variable resistor in such a way that the galvanometer needle rests at zero in the absence of alternating voltage. When the zero-setting procedure is performed, the input circuit of the voltmeter should be short-circuited to prevent the instrument from picking up stray induced a.c. voltage. Fig. 261*d* shows the circuit of a valve voltmeter employing a double-diode type 6X6C and incorporating a compensation of the initial anode current. The scale of a diode valve voltmeter is always somewhat compressed in the beginning, which is attributed to the effect of the lower bend of the valve characteristic.

Some semiconductor diodes, for example germanium diodes, may be successfully used in the place of electron valve diodes. The advantages of electronic voltmeters employing such semiconductor diodes are seen in that a voltmeter of this type needs no filament supply, and, hence, is free of the initial current effect. Because of the absence of such effect, the semiconductor voltmeter requires no initial current compensation.

Dealing with valve voltmeters, we have so far been speaking of measuring instruments using only diode valves. However, some valve voltmeters are built around triode valves and possess a definite advantage over the diode voltmeters in that they have a much higher input impedance. The shortcoming of the triode voltmeter is that the device requires a source of anode voltage. In most cases, a triode valve voltmeter is nothing but an anode detector stage in which the d.c. component I_{a-} is the object of measurement. The value of this component increases when alternating voltage is applied to the grid of the valve. Thus, the anode galvanometer can be calibrated in terms of the alternating voltage being measured. The scale divisions are not of equal size in the triode voltmeter, just as in the case of the simpler diode voltmeter.

The anode detector is always characterised by the presence of a certain initial current I_{a0}. This current is compensated for by passing a certain amount of auxiliary current through the galvanometer from the anode power supply, such compensating current being equal to the initial current but opposite to it in direction. The bridge circuit of compensation (Fig. 262*a*) offers the maximum convenience in this respect. In this circuit resistors R_1, R_2, R_3 and the anode resistance of the valve (together with R_c) form the bridge. The galvanometer is connected into one diagonal of the bridge, the anode power supply—into the other. When the bridge is balanced, no current flows through the galvanometer. When a.c. voltage is applied to the grid of the valve, the balance of the bridge is upset, as far as d.c. current is concerned, and the galvanometer registers a deflection. The circuit shown in the drawing has a closed-circuit input. The value of resistor R_g must be at least several megohms, preferably from 10 to 50 megohms, in order to make the input resistance sufficiently great. The

value of isolating capacitor C_g may be comparatively low. In order to obtain the anode detection, the necessary bias voltage is provided by cathode resistor R_c. Capacitor C bypasses the a.c. component of the anode current directly to the cathode, keeping it away from the resistor. One of the resistors R_1, R_2, R_3 or R_c is made variable. This is done to ensure the preliminary balancing of the bridge and the setting of the galvanometer to zero.

The anode detection tolerates no grid currents. Because of this, the amplitude of a.c. voltage being measured must not exceed the value of the bias voltage, i.e., it should not exceed a few volts, as a rule. When greater measurement ranges are required, it becomes necessary to provide a voltage divider (Fig. 262*b*), made up of several high resistances, at the input of the voltmeter circuit. However, when dealing with high frequencies, such a divider does not give equal

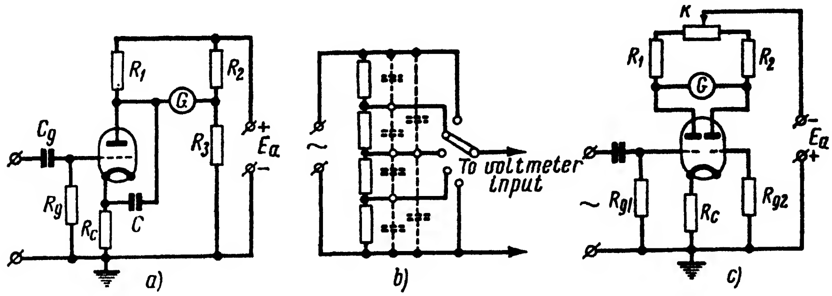


Fig. 262. Bridge-type triode valve voltmeters

voltage distribution on various frequencies. This effect is attributed to the influence of parasitic capacitances shunting separate sections of the divider (the parasitic capacitances are indicated by broken lines in the drawing). It is, therefore, preferable to increase the value of the cathode resistor (which accounts for the negative feedback) when performing high-voltage high-frequency measurements with the type of voltmeter being described. This can be appreciated from the following. When the a.c. voltage fed to the grid of the valve is increased, the direct current component I_{a-} increases, which is followed by an increase of grid bias voltage developed across the automatic biasing resistor. As a result, it becomes possible to measure higher voltages without running into the grid current region. Besides this, the voltmeter scale becomes nearly equally-spaced when the value of R_c is made greater.

The triode-type valve voltmeter may be used for d.c. voltage measurements, if desired. In this case, the positive side of the voltage being measured should be connected directly to the grid of the valve. The scale calibration on d.c. measurements will vary from that on a.c. measurements. If the negative side of the d.c. voltage being measured is connected to the grid, the initial negative bias voltage should be set to a value not exceeding —1 v. When this is done, the current flowing through the galvanometer will be reversed and polarity of the galvanometer will have to be changed. Any type of rectifier may be used as the power supply for the described type of voltmeter and there is no need to provide a thorough smoothing at the output of such a rectifier. When the voltmeter is first switched on, the cathode of the valve will take a little time to reach the normal operating temperature. During this period of time, the bridge is very unbalanced and the current flowing through the galvanometer can be excessively high. In order to protect the galvanometer from damage, in the circuit given in Fig. 262*a* the anode voltage may be applied only after the cathode has reached its normal working temperature. The circuit given in Fig. 262*c* is free from this shortcoming, because it uses an auxiliary valve in place of the resistor R_3. In this circuit, the cathodes of the two valves are heated simultaneously, which

precludes a great unbalance of the bridge. Potentiometer R is employed for balancing the bridge, i.e., for setting the galvanometer pointer to zero. A circuit of this type also has another advantage; the bridge balance is not upset even when the operating condition of the power supply varies within certain limits. This is explained by the fact that the anode currents of both valves change to the same extent during power supply voltage variations. The greater the similarity of the characteristics and parameters of the two valves, the lower is the sensitivity of the circuit to the variations of the power supply voltages.

Of great interest is the circuit of a valve voltmeter shown in Fig. 263. This circuit is widely used and, in particular, is incorporated in valve voltmeters type BKC-7 of Soviet manufacture. In this circuit, the first stage is represented by an amplitude diode voltmeter with closed input arrangement. The value of resistor R reaches several dozens of megohms. With such a high resistance in the circuit, there is no need to employ too great a value of C even on the lowest frequencies

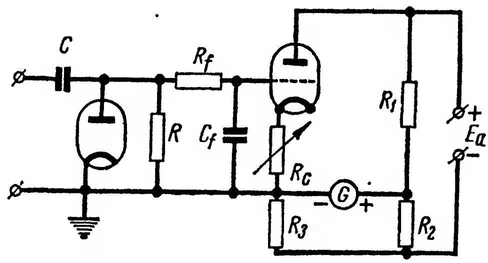


Fig. 263. A two-stage valve voltmeter

being measured. A triode or a more complex valve, connected as a diode, is frequently employed in this input stage of the voltmeter. When the voltmeter is to measure a.c. voltages in the ultra-high frequency range, an acorn or a bantam valve should be used in this stage.

In the voltmeter being described, d.c. voltage appears across resistor R, the value of this voltage being nearly equal to the amplitude value of the a.c. voltage under measurement. The negative side of this d.c. voltage is fed to the grid of the valve in the second stage. Filter R_fC_f blocks the path of the a.c. component of the rectified voltage, keeping this voltage away from the grid of the valve. The value of R_f is about 10-20 megohms, that of C_f —several thousand picofarads. The second stage forms a bridge circuit. Here, resistors R_1, R_2, R_3 and the anode resistance of the triode (together with R_c) represent the arms of the bridge. The compensation of the initial anode current is performed in this stage. If the given stage employs a low mutual-conductance triode, for instance, type 6C5, the scale of the galvanometer used with the voltmeter should be calibrated from zero to about 0.5 ma. Using a high mutual-conductance valve (for instance, beam tetrode 30П1С, connected as a triode and having $S = 10$ ma per volt) calls for a milliammeter of 2-3 ma.

Cathode resistor R_c serves to create a small initial bias, making the stage operate without grid current. When the input terminals of the voltmeter are short-circuited, resistor R is adjusted to obtain the balance of the bridge, as indicated by the zero position of indicating meter. If an a.c. voltage is applied to the input terminals of the voltmeter, a considerable negative bias voltage is developed across resistor R and is applied to the grid of the second stage valve. This bias shifts the operating point to the left on the grid characteristic, the bridge balance becomes upset, and the galvanometer gives a reading. At a certain value of the voltage being measured, the anode current of the second stage valve is cut off. This means that the bridge balance has been upset as much as possible. Under such condition, the galvanometer pointer deflects to give the maximum reading. If the voltage at the input terminals of the voltmeter is

increased further, no change in meter reading will result, because the triode is already at its cut-off condition. Thus, the voltmeter being described cannot be damaged when an excessive voltage is applied to its input. If the instrument is to be provided with several scales, resistance R may be replaced by a voltage divider, a rotary switch cutting in various sections of such a divider into the circuit. This permits to apply to the grid of the second valve only a part of the voltage rectified by the first valve. However, when the voltmeter is designed to measure high-frequency voltages, the voltage divider arrangement must not be used at the input of the meter.

This voltmeter may be easily arranged to measure d.c. voltages. In this case, the voltage must be applied to R in such a way that the negative side of the voltage is fed to the grid of the second valve. It is not necessary to disconnect the diode valve. The d.c. calibration almost exactly coincides with the calibration related to the alternating-voltage amplitudes.

None of the various types of voltmeters discussed above are suitable for the measurement of very low a.c. voltages ranging from a few millivolts to a few dozens of millivolts. When such low voltages are to be measured, a preliminary multi-stage amplifier must be incorporated in the voltmeter circuit. The gain of such an amplifier must be stabilised by means of negative feedback.

It is highly desirable that the power supply used with a valve voltmeter has a high order of voltage stability. For this purpose rectifier power supplies employed by such voltmeters frequently incorporate automatic voltage stabilisation circuits. When a rectifier is not provided with such a circuit, voltage stabilisation may be effected in a simpler way — manually. In such a case, a rheostat is connected in series with the primary winding of the power transformer. Connecting the voltmeter itself to measure the filament voltage of its valves and adjusting the rheostat to obtain the rated filament voltage reading offers a method of setting the correct operating condition of the valve voltmeter by hand.

Application of Neon Lamps

Sometimes, neon lamps may be used as indicators of d.c. voltage and also of a.c. voltage of any frequency. As we already know from the previous studies, a neon lamp is characterised by the so-called firing or ignition voltage. This is the voltage at which the neon lamp suddenly begins to conduct current and to glow. Indicating this voltage as U_f, we note that U_f is quite definite in d.c. circuits and has a certain predetermined value for each type of neon lamp. (Chapter IV, Sec. 51.)

If an alternating or a pulsating voltage is applied to a neon lamp, the lamp will glow, provided that the amplitude (the maximum) value of the voltage is not lower than U_f. Owing to this property of a neon lamp, it is possible to build very simple neon instruments for the purpose of approximate measurement of voltage. Fig. 264 shows an instrument of this kind. The instrument is designed to measure d.c. voltages and also the amplitude values of low-frequency a.c. voltages. The voltage to be measured is applied across a voltage divider consisting of two resistors R_1 and R_2. The neon lamp is connected in parallel with R_2. To take a measurement of voltage, the value of resistor R_1 is slowly decreased until the lamp suddenly flashes on. Prior to the measurement, the circuit is first calibrated and the calibration is marked off along the scale of variable resistor R_1.

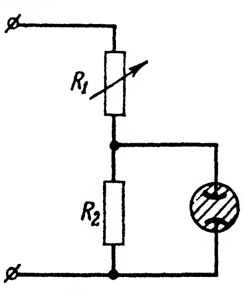


Fig. 264. A neon-lamp voltage measuring device

Of course, the accuracy of such voltage measurement method is not high because U_f varies with the ambient temperature and with the ageing of the neon

lamp. Besides, the method described makes it possible to measure only such voltages which are higher than U_f. True, voltages lower than U_f could also be measured with the help of a neon lamp, but this calls for a separate step-up transformer with a well-known transformation ratio.

This primitive measuring circuit can also be used for approximate determination of maximum values of pulsating voltage. If it is required to determine the amplitude of the a.c. component of such a voltage, a blocking capacitor of sufficient capacitance must be included in the circuit. Such neon lamp devices are not suitable for high-frequency voltage measurements because resistor R_2 is shunted by the interelectrode capacitance of the lamp, the reactance of such capacitance varying on different frequencies. Still, high-frequency voltages may be measured with the neon device, if resistors R_1 and R_2 are substituted by capacitors. The considerable, and varying, input capacitance of the neon device is its main shortcoming, making it hardly suitable for high-frequency measurements, when any kind of accuracy is desired. Generally speaking, the connection of any device, possessing a certain input capacitance, to a tuned circuit, for the purpose of high-frequency measurement, always upsets the resonant condition of the circuit and makes it necessary to restore the resonance by decreasing the capacitance of the tuned circuit capacitor.

Calibration of Voltmeters

Voltmeters designed for operation in direct-current circuits are, naturally, calibrated with the help of a d.c. power supply. Rectifier-type and valve voltmeters are usually calibrated with the help of 50-cps mains voltage. It is desirable to employ a source of high-frequency when calibrating valve voltmeters designed for high-frequency measurements. This, however, is not always possible, because a standard high-frequency voltmeter is often difficult to obtain. On the other hand, when the calibration is performed on the usual industrial frequency of 50 cps, a good moving-coil voltmeter can be used as a reference instrument.

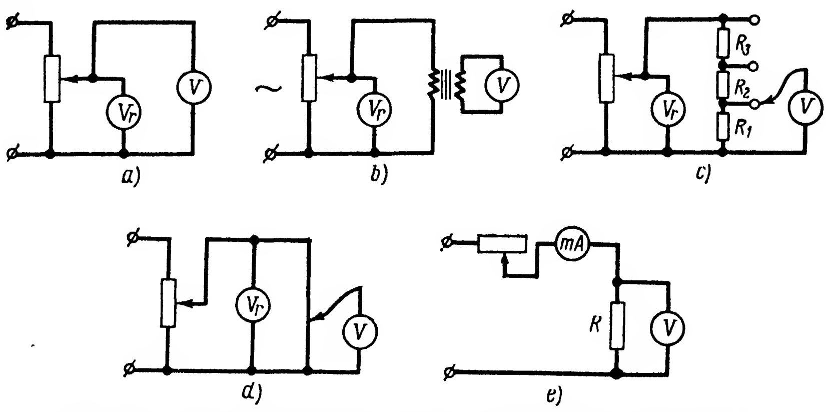


Fig. 265. Circuits employed for the calibration of voltmeters

Fig. 265 gives several types of calibration circuits. Circuit shown in Fig. 265*a* is the one most frequently used. In this circuit a reference voltmeter V_r and voltmeter being calibrated (V) are connected in parallel, a potentiometer providing the voltage adjustment. If the instrument being calibrated is designed for a lower voltage than the voltage of the power supply (rectifier, mains), the potentiometer must be supplied with power through an additional dropping

resistor or a step-down transformer. It may also happen that the reference voltmeter is designed for a higher or lower voltage than that required for the calibration. If such be the case, one of the two voltmeters is connected to the other through a step-down transformer with a definite transformation ratio, or, alternatively, through a voltage divider, the voltage division ratio of which is exactly known. These two alternative methods are shown in Fig. 265 *b* and *c*, respectively.

When the voltage divider scheme of calibration is employed, the following should be observed. The resistance of that part of the voltage divider which is connected in parallel with the voltmeter under test (R_1 in Fig. 265*c*) must be may times lower than the input resistance of the voltmeter itself. This precaution is necessary in order to reduce the shunting influence of the voltmeter as much as possible. This precaution is not required for that voltmeter which is connected across the whole of the voltage divider; thus the reference voltmeter V_r in Fig. 265*c* may be a low-resistance type of instrument.

Fig. 265*d* shows a circuit devised specially for the calibration of very low-scale voltmeters measuring voltages as low as fractions of one volt. In this circuit, the calibration voltage is applied across a thin high-resistance wire, 1 metre long, while the voltmeter being calibrated is connected to various points of such wire. Each centimetre of the wire length corresponds to 0.01 part of the voltage read by the reference voltmeter.

In some cases, an ammeter or a milliammeter may be used as a reference instrument. The current passing through such a reference meter is made to flow through the resistor *R* (Fig. 265*e*). The exact resistance value of *R* is known and, therefore, the voltage drop across this resistor may be readily calculated. The meter to be calibrated is connected across *R* and the current value is adjusted by means of a rheostat. If the voltage must be lowered, a voltage divider has to be connected between *R* and the voltmeter.

122. RESISTANCE MEASUREMENTS

The Method of Voltmeter and Ammeter

When this method of resistance measurement is resorted to, two instruments are required—a voltmeter and an ammeter (Fig. 266 *a* and *b*). The resistance is determined from Ohm's law. The circuit given in Fig. 266*a* is used to measure

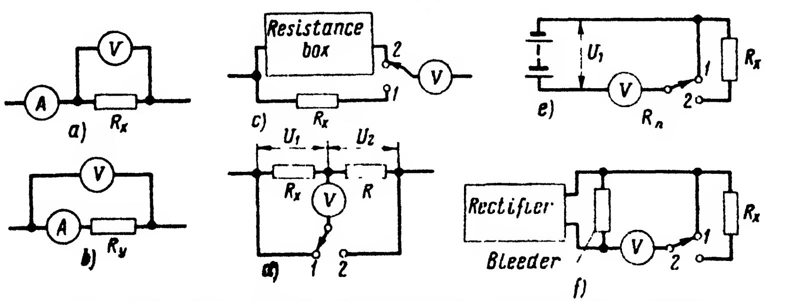


Fig. 266. Direct-current resistance measurement circuits

such resistances whose values are many times smaller than the resistance value of the voltmeter, while the circuit shown in Fig. 266*b* is suitable for measuring resistances whose values are considerably greater than the resistance value of the ammeter (or milliammeter). If the voltage of the power supply source is known and its internal resistance is low, the voltmeter can be dispensed with and only the ammeter used for the measurement.

The Substitution Method

In this method, any type of available meter (galvanometer, milliammeter or voltmeter) is connected in series alternatively with the resistance being measured (R_x) and a resistance box (Fig. 266*c*). The resistance box is so adjusted that the meter gives the same reading whether connected to the box or to resistance R_x. When such a condition is attained, the value of R_x is determined from the setting of the resistance box dials. The described method of resistance measurement offers a high degree of accuracy and does not require a special exactly-calibrated reference instrument, but it does require a resistance box.

The Comparison Method

Here, the resistance to be measured (R_x) is connected in series with a known resistance R (Fig. 266*d*). A voltmeter, connected as shown in the circuit diagram and possessing a much higher internal resistance than R_x and R, is used to measure the voltages built up across these two resistors. Denoting these two voltages, respectively, as U_1 and U_2, the value of R_x may be computed from the following relation:

$$\frac{R_x}{R} = \frac{U_1}{U_2},$$

or:

$$R_x = R\,\frac{U_1}{U_2}.$$

The smaller the difference between the resistance values of R and R_x, the more accurate will be the result of the measurement.

The Method of Voltmeter

The method of measurement here described (Fig. 266*e*) is the simplest of all resistance measurement methods, although it does not provide high accuracy. This method requires but one measuring instrument—a voltmeter, whose internal resistance R_n must be exactly known beforehand. If unknown, it has to be measured first. Some voltmeters carry a designation of the current consumed by them to provide a full-scale deflection. In such a case, the internal resistance of the voltmeter can be readily determined by taking the voltage corresponding to the full-scale deflection and by dividing this voltage by the consumption current marked on the instrument.

The procedure of resistance measurement by this method is as follows. First, the voltmeter is used to measure voltage U_1 of the power supply. After this, the unknown resistance R_x is cut into the circuit in series with the voltmeter (see diagram in Fig. 266*e*). This done, the voltmeter will give another reading, which may be referred to as U_2. U_2 will be smaller than U_1. Having recorded both of these readings, the value of R_x may be found from the following formula:

$$R_x = R_n\left(\frac{U_1}{U_2} - 1\right).$$

The above method will give the greatest accuracy if it happens that $R_x = R_n$. This method should not be resorted to when the value of the resistance to be measured is known to be smaller than 0.1 R_n or greater than 10 R_n. The power supply source used in this measurement should have a low internal resistance; this will ensure a constancy of voltage U_1 when the resistance of the circuit is changed while R_x is being cut in and out of the circuit.

If a power supply source with a considerable internal resistance (for instance, a rectifier) has to be used in the measurement, a low-resistance bleeder resistor must be connected across the output terminals of such a power supply (Fig. 266*f*). If the resistance of the bleeder is much smaller than R_n, the voltage supplied by the power source will remain practically unchanged during the described switching operation.

Ohmmeters

Fig. 267*a* shows the series circuit of an ohmmeter. The main components of this instrument are: a milliammeter, power supply (a dry cell or a small dry-cell battery), and series resistor R. When nothing is connected to the instrument terminals (to which the resistance to be measured will be later connected),

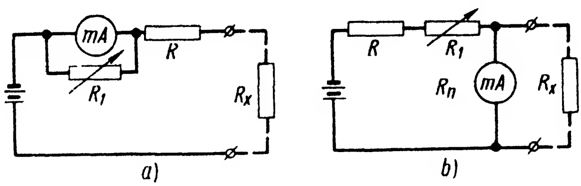


Fig. 267. Ohmmeters with series *(a)* and parallel *(b)* circuits

no current flows through the milliammeter and the pointer of this meter is positioned at the beginning of the scale, pointing at mark ∞. The latter mark indicates that an infinitely high resistance is connected across the terminals of the instrument ($R_x = \infty$ is the same thing as R_x taken out of the circuit altogether).

When the instrument terminals are short-circuited, the current flowing through the meter reaches its maximum value. Under such condition, the meter pointer will give a full-scale deflection, swinging to the end position of the scale, this position marked by zero (0).

The short circuit removed, and the unknown resistor R_x connected across the instrument terminals, the value of current flowing through the meter will be determined by the resistance value of R_x. Such resistance value of the externally-connected resistor will correspond to a certain and definite position of the meter pointer. This is the reason why the scale of the instrument being described can be directly calibrated in ohms. The divisions of such a scale are not equally spaced; they are compressed in the region of high resistance values. A sufficient accuracy of measurement is obtained if the value of the unknown resistor R_x is within the limits of 0.1-10 times the resistance value of R. The circuit of such an ohmmeter includes a variable resistor, whose value is 10-20 times greater than the resistance value of the measuring instrument. This resistor, marked as R_1 in the circuit diagram, is adjusted to vary the sensitivity of the ohmmeter within certain limits, and also to compensate for the gradual decrease of the battery voltage, as the service life of the battery approaches its natural end. The procedure of setting the ohmmeter to its normal sensitivity is as follows. The instrument terminals are first short-circuited. After this the knob of resistor R_1 is rotated until the meter pointer is brought to the zero setting, whereupon the instrument is ready for use. The described procedure should be carried out every time before the ohmmeter is put into operation; this will assure that the instrument operates at its normal sensitivity in every case.

There is an alternative way of varying the sensitivity of the meter; some ohmmeters dispense with the described variable electrical shunt R_1 and employ

a special magnetic shunt in its place. The magnetic shunt is nothing but a steel plate located between the pole pieces of the milliammeter magnet. A part of the magnetic flux is made to flow through this steel plate, and when the latter is shifted, the meter sensitivity varies. Since the meter pointer is positioned at infinity (∞) when no current flows through it, the usual adjustment screw, provided in all meters, is used to bring the pointer to this required position of ∞. Prior to putting the ohmmeter into operation, the infinity setting is first checked and, if required, corrected by means of the adjustment screw. Following this, the zero setting is made with the help of the magnetic shunt knob. If the meter pointer refuses to be set to zero, this is an indication that the battery is run down. When it is required to measure high unknown resistances (R_x), both the value of R and the value of the battery voltage should be increased.

Low resistances are sometimes measured with the help of the circuit shown in Fig. 267*b* and known as the parallel ohmmeter circuit. In this circuit, the unknown resistor R_x is connected in parallel with the milliammeter, while series resistance R must be considerably greater than resistance R_n of the meter itself. Variable resistor R_1 performs the infinity adjustment (the instrument terminals must be disconnected from the unknown resistor R_x during this adjustment). In this type of ohmmeter, the zero is located at the left and the infinity mark at the right of the instrument scale. The measurement limits lie approximately between $0.1R_n$ and R_n.

Very great resistance values are measured with the help of special instruments called *meggers*. The megger usually employs a small hand-driven generator as a source of voltage, the generator developing 100-200 volts and sometimes higher. Alternative types of meggers are powered from rectifiers and from vibra-packs.

When a megger is not available, very high resistances (dozens of megohms and higher) may be measured by connecting them in series with a microammeter and any type of high-voltage source developing 100-300 volts. The procedure, in this case, is as follows. If the power supply voltage is equal to, say, 200 v, and the microammeter (for instance, of the 100-μa type) gives a reading of 8μa, the value of the unknown resistance is given as:

$$R_x = \frac{200}{8\times10^{-6}} = 25\times10^{6} \text{ ohms} = 25 \text{ megohms}.$$

To prevent the microammeter from an accidental burn-out in the described measurement, it is recommended to connect into the circuit a protective series resistor, whose value is equal to the voltage of the power supply divided by current corresponding to the full-scale deflection of the meter pointer. In the example given above, the value of such resistance is as follow:

$$R = \frac{200}{0.1\times10^{-3}} = 2{,}000{,}000 \text{ ohms} = 2 \text{ megohms},$$

and then the true value of resistor R_x is equal to 23 megohms, when a current of 8μa flows through the circuit.

The Bridge Method of Resistance Measurement

Fig. 268 gives several bridge circuits employed for resistance measurements. The bridge measurement principle is based on the equilibrium of bridge arms, i.e., on the absence of current in the diagonal AB of the bridge when the following relation is established between the bridge arms: $R_xR_2 = R_1R_3$. In other words, the condition of equilibrium (otherwise called the condition of balance) sets in when the resistance products of the opposite arms are equal. Hence:

$$R_x = \frac{R_1}{R_2} R_3.$$

In the circuit of Fig. 268a, resistance R_3 is the standard resistance, usually called the reference resistance. In the same circuit, resistors R_1 and R_2 may be represented by a straight piece of thin rheostat wire or by a potentiometer. For the purpose of obtaining the condition of balance, the ratio $R_1 : R_2$ is changed by means of slider A. In the circuit of Fig. 268b, R_1 and R_2 are the reference resistors, by means of which various ratios may be set up (for instance 100:1; 10:1; 1:1; 1:10; 1:100, etc.). In the same circuit, variable resistor R_3 is adjusted to secure the balance of the bridge.

In the bridge circuit (Fig. 268a), a sufficient accuracy of measurement is obtained when the ratio $R_1 : R_2$ lies within the limits of 0.1-10. The maximum accuracy is secured when $R_1 : R_2 = 1$. In general, when high accuracy is sought, the resistance of all the arms in a bridge should be approximately equal.

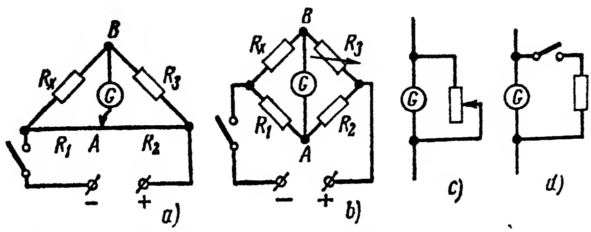


Fig. 268. Bridge circuits for resistance measurements

When a bridge is supplied with power from a d.c. source, the galvanometer used as the current indicator and connected into the bridge diagonal should have the zero position in the middle of its scale. Since the bridge is usually way out of balance when a measurement begins, excessive current can flow through the galvanometer, damaging this meter. Because of this, means should be provided to reduce the galvanometer sensitivity at the beginning of each measurement. The protection of the galvanometer is usually attained by shunting the meter by a variable resistor or a fixed resistor with a switch (Fig. 268 c and d). As soon as an approximate balance has been found, the shunting resistor is disconnected from the meter, after which a more precise balance is found.

In some cases, the bridge is fed with an alternating current of a frequency of a few hundred cps. The current may be generated by an audio oscillator or a buzzer. An earphone is used in place of the galvanometer in the a.c. bridge. The bridge is adjusted until the earphone no longer reproduces the audio note. This is a sign that the condition of bridge balance has been reached.

Bridge circuits powered from a d.c. supply sources can also employ an earphone as a balance indicator. In this case, the condition of balance is evidenced by the disappearance or by the maximum weakening of clicks in the earphone as the power supply switch is thrown on and off.

Circuit Checkers

Circuit checkers are used when only a rough check of a circuit or its components is desired (for instance when looking for a break, for a short-circuited capacitor, etc). Fig. 269 shows several types of such circuit checkers.

Terminals (or simply the leads) a and b of the checker are connected to the circuit being checked. The device using a lamp requires a freshly-charged battery in order to make the lamp give a good glow if the circuit being checked is not open. This checker is suitable for testing only such circuits whose resistance exceeds the resistance of the check lamp not by more than a few times. All the

other circuit checkers shown in Fig. 269 can give satisfactory performance even when the battery is considerably discharged.

The performance of the voltmeter-type and milliammeter-type checkers is about the same. Resistor R, connected in series with the milliammeter, is so selected that the meter gives a full-scale deflection when the checker terminals a and b are short-circuited. If the value of resistor R is known (and if the battery voltage is anywhere its normal value—*Translator's note)*, the reading obtained on the milliameter will give a rough idea of the resistance value of the circuit being checked; in this case the circuit checker serves as a primitive ohmmeter. If the voltage of the checker battery is small, the device cannot be used to check high-resistance circuits, because the meter simply will not give any reading. For instance, if the voltage of the battery is equal to 4 v and a milliammeter with the scale of 10 ma is employed, the value of the series resistor used with

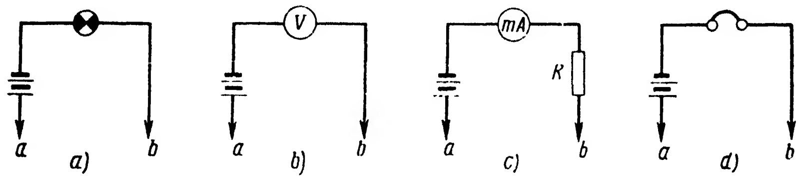


Fig. 269. Various types of circuit checkers employing: an incandescent lamp *(a)*; a voltmeter *(b)*; a milliammeter *(c)*; an earphone *(d)*

this milliammeter is equal to: $R = \frac{4}{0.01} = 400$ ohms; then, when checking a 1-megohm circuit, the current will be equal to: $\frac{4}{1,000,000} = 4 \times 10^{-6}$ a $= 4$ microamperes. Of course, in such a case, the milliammeter pointer will remain at zero.

The circuit checker with the earphone is the most sensitive of all the checkers listed above, because the given checker gives a fairly good performance even when the current flowing through it is as low as a fraction of a microampere. When this checker is connected to a circuit whose resistance is as high as several megohms, the device will—even in such a case—register the circuit continuity. The continuity of the circuit is indicated by the clicks obtained when the ckecker circuit is closed and opened. This means that the circuit is not disrupted, so far as the direct current is concerned. The value of resistance in the circuit may be only very approximately determined by the intensity of the clicks. If the circuit is disrupted, only a slight click will be heard on closing of the contact; such a click is attributed to the initial passage of current through the capacitance of the circuit being checked. For instance, if the circuit checker is used to check the continuity of a transformer winding and a weak click is heard when the circuit is closed but not when it is opened, this is an indication that there is a break in the winding. The two disrupted parts of such a winding form a capacitor which will take a certain amount of current to charge up at the moment the circuit is closed. A similar effect is experienced when the checker is used to test an iron-core coil for the absence of a short circuit. If the checker is employed to test a capacitor, a click when the circuit is closed and the absence of the click when the circuit is opened is an indication that the capacitor is not faulty, i.e., not punctured and does not have a considerable leakage. The greater the capacitance of a good capacitor being checked, the louder will be the click on closing of the circuit.

The effect described above is also experienced with the checker supplied with a meter; the meter, of course, can reproduce no clicks, but it will register a reading only at the moment its circuit is closed when testing a good capacitor, the meter pointer returning to the zero position soon after the closing of the cir-

cuit. It should be noted, however, that when the meter-type checker is conneced to a low-capacitance capacitor, the charging current will be so small that the meter will give no deflection. The meter-tester has a smaller sensitivity than the telephone-tester.

A circuit tester will show the presence of leakage in a capacitor; the presence of leakage is determined by the time the capacitor holds its charge. This test is performed by first charging the capacitor and then immediately discharging it through an earphone or a meter, recording the deflection of the meter pointer or the loudness of the click. Then the capacitor is charged again, to be discharged some dozens of seconds later. During this delayed discharge the presence of leakage can be discovered; the larger the leakage, the smaller will be the deflection of the meter pointer or the weaker will be the click in the earphone.

As already stated, low-capacitance capacitors do not provide a noticeable meter pointer deflection. Hence such capacitors should be tested only with the earphone-type ckecker. When capacitors of the electrolytic type are tested, the delayed discharge is tested after only a few seconds, because such capacitors normally possess high leakage, and waiting for dozens of seconds will weaken the discharge considerably.

When testing components of any circuit care should be taken that other components, connected into the same circuit, take no part in causing the meter deflection or the click in the earphones. Should such be the case, the offending components must be disconnected from the circuit under test. For instance, when checking the leakage of a capacitor shunted by a resistor, the shunt must be disconnected before making the test. Before applying the circuit checker to a resistor, make sure that no other resistors are connected in parallel with such a resistor. If such shunting resistors are present, they must also be disconnected, unless their resistance value is much higher than the resistance value of the resistor to be tested. It is, of course, to be understood that when the components being tested are a part of a circuit, the power supply of the circuit should be switched off before the test can be made.

123. AUDIO-FREQUENCY OSCILLATORS

Audio-frequency valve oscillators are widely used for various types of measurements, as well as for testing amplifiers, receivers and transmitters. In order to provide a high degree of performance in such applications, the oscillators of this type must possess very stable frequency, amplitude and calibration. Besides this, the oscillators must provide a sinusoidal output voltage free of harmonic content, and give the necessary power output and output voltage, both of them being capable of easy and convenient adjustment. And, of course, such oscillators must possess an audio range sufficiently wide for various types of measurements. Some test oscillators of this type are provided with a rectifier-type, or valve, voltmeter to measure the output voltage generated by the oscillator. It is highly desirable that the circuit and construction of such oscillators are as simple as possible. The oscillators should be economical in the power they draw from the supply source, the latter usually being a rectifier. If best measurement results are to be obtained from a test oscillator, the voltage of the supply source should be stabilised.

Below is given the description of principal types of audio-frequency oscillators.

LC audio-frequency oscillators. The oscillators of this type directly generate audio-frequency power without any intermediate frequency conversion circuits. Such oscillators employ tuned circuits comprised of inductance coils and capacitors, the circuits adjusted to various frequencies within the audio-frequency range. It should be noted, however, that it is difficult to cover this entire range by means of such circuits. The range extends from 20 to 20,000 cps, which means that the ratio of the highest to the lowest frequency must be 1,000. Apparently,

the LC product has to be changed within the limits of 1,000,000 times in order to attain the required frequency coverage!

This is a very difficult assignment. The job can be done only when the entire audio-frequency range is split up into many frequency bands. Moreover, in the lowest frequency band, the required frequencies would have to be obtained with the help of very high inductances. Such inductances can be provided only by coils of extremely large size, wound with a great number of turns, supplied with ferromagnetic cores, and noted for considerable non-linear distortion introduced by them into the associated circuit.

Because of all these difficulties, LC audio-frequency oscillators are generally designed to generate alternating test voltage at one or, at most, at a few fixed frequencies. For instance, the simplest tests of radio receivers and amplifiers are carried out with the help of a specially-designed LC test oscillator operating at a single frequency of 400 cps, this frequency being considered the average frequency in such tests. Oscillators of this type usually employ a single valve and an inductive or autotransformer feedback arrangement. On such a low frequency even a self-excited oscillator can provide a sufficiently good frequency stability.

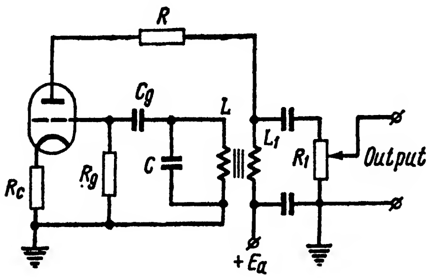


Fig. 270. The simplest type of audio-frequency oscillator

Fig. 270 gives an example of the simplest circuit of the audio-frequency test oscillator discussed above. An interstage transformer or any other low-frequency transformer may be used as L and L_1. Capacitor C, shunting the primary winding of the transformer, is so selected that the oscillator generates the required frequency. The magnitude of the output voltage generated by the circuit is adjusted by means of the 0.5-megohm potentiometer R_1. If the inductance of the transformer windings is sufficiently high, the lowest frequency generated by the oscillator can be obtained even without including capacitor C into the circuit. In this case, the distributed capacitance of the transformer takes the place of the capacitor C. When it is desired to obtain several frequencies from the described type of oscillator, the coils are tapped and appropriate taps are connected into the circuit by means of a rotary switch. An alternative way of doing the job is the employment of the rotary switch and of several different capacitors. The inclusion of grid resistor R_g into the circuit of the oscillator cuts down the energy consumption of the stage and also helps to stabilise the amplitude of oscillations. The shape of the oscillations is improved by connecting resistor R into the anode circuit, the value of such resistor made equal to several tens of thousands or several hundreds of thousands of ohms. The greater the value of such resistor, the smaller will be the harmonic content in the output of the oscillator although the smaller will also be the output power of the stage. If the value of R is made too high, the oscillator will cease functioning. A considerable negative feedback value will also provide a high attenuation of harmonics. Such feedback is obtained by the inclusion of a high resistance in the cathode lead of the oscillator. The value of the resistor should be between several hundred or some thousand ohms, and the resistor must not be shunted by a capacitor. In the oscillator of this type, the value of the output load impedance exerts a considerable influence upon the frequency generated by the oscillator. Thus a considerable change of frequency is experienced when the potentiometer R_1 is varied to adjust the output voltage of the oscillator. The frequency, in this case, may be stabilised by including an amplifier stage, following the oscillator. An alternative way of stabilising the frequency is the employment of an electron-coupled or a transitron-type oscillator stage.

Beat-frequency oscillators. The beat-frequency oscillator is more complex than the *LC* oscillator described above. Moreover, it has a poorer frequency stability, although it employs several electron valves. Still, this type of oscillator has an important advantage over the *LC* oscillator. This advantage is the coverage of the entire audio-frequency range without any need for switching and without the necessity of employing the clumsy tuned audio-frequency circuits. The circuit of the beat-frequency oscillator is given in Fig. 271*a*.

Two high-frequency heterodynes make up the circuit of the beat-frequency oscillator. One of these oscillators, denoted as H_I in the circuit diagram, has a fixed frequency. The other oscillator, denoted as H_2, generates a voltage of frequency variable over a certain narrow range.

The frequencies of the two heterodynes are usually selected within the range of 100-200 kc. As an example, the oscillators shown in Fig. 271*a* are assumed to

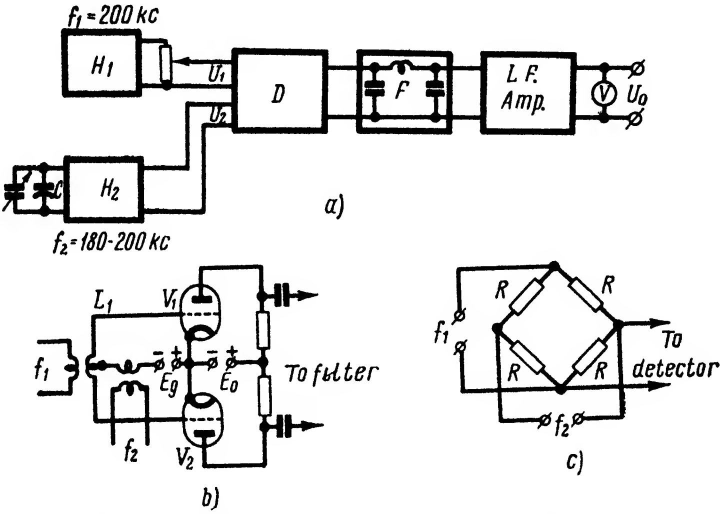


Fig. 271. Beat-frequency audio-oscillator circuits

operate, respectively, on frequencies: $f_1 = 200$ kc and $f_2 = 180$-200 kc. The oscillations developed by both heterodynes are fed to detector *D*. This is a converter stage where the two oscillations beat with each other and produce a resulting beat frequency. As a result of this, the difference frequency $F = f_1 - f_2$ appears at the output of the detector. Frequency f_1 normally kept constant, the value of f_2 is varied to produce beat oscillations whose frequency F ranges from 0 to 20 kc. This is obtained by changing f_2 only by 10%, which is easily accomplished. Thus, the whole audio-frequency range is covered by a single rotation of the rotor of the tuning capacitor belonging to heterodyne H_2.

The beat frequency signal developed in the detector *D* is fed to a low-frequency amplifier for the purpose of amplification. Filter *F*, comprised of high-frequency chokes and capacitors, is provided at the output of the detector to suppress the high-frequency oscillations and keep them away from the low-frequency amplifier. A greater stability of the resultant beat frequency (audio frequency) is secured when the two heterodynes are similarly designed. This done, the frequency drift of the heterodynes, such as may be caused by external factors, will be about the same. As a result, the difference frequency will remain constant. It is regrettable, however, that the two heterodynes H_1 and H_2 cannot be made quite the same, because one of them is intended for fixed-frequency operaton, while the other — for variable-frequency performance. In view

of the above, the tuned circuits of the two heterodynes have to be of different design. Because of this, the inevitable frequency changes of the two heterodynes will not be similar, which will cause a certain instability of the resulting audio frequency. Such instability is particularly pronounced on low audio frequencies, when f_1 and f_2 are close to each other. This may be seen from the following. Assume that, as a result of heating of the circuit elements, f_1 changes by 100 cps, while the change of f_2 is equal to 110 cps. Evidently, when this happens, F will change by 10 cps. The 10-cps change will not be noticeable when the difference frequency F is expressed in several thousand cycles per second. However, if $F = 50$ cps, the 10-cps change is equivalent to 20%, which is quite impermissible. The described effect is the main shortcoming of the beat-frequency oscillators.

Frequency instability upsets the correct calibration of the instrument, but can be compensated for by including trimming capacitor C into the tuned circuit of one of the two heterodynes. The calibration is then set right by the method of zero beats. This is done by setting the scale of the oscillator to $F = 0$ cps and by adjusting the compensating trimmer C in such a way that actually zero frequency is obtained. The moment when F becomes equal to zero is determined by a corresponding reading of the output voltmeter or of an electron-beam indicator. When F is equal to a few cps, the voltmeter pointer, or the glowing part of the indicator screen, will register an oscillation at the given frequency, the effect being quite discernible by the eye. The oscillation becomes slower and slower as frequency F approaches zero, and finally stops as the exact zero-beat condition is reached.

In some cases, the calibration correction is carried out at the industrial frequency of 50 cps. To do this, a voltage of the frequency of a.c. mains (taken off, for example, from the filament winding of the power transformer) is applied to the input of the low-frequency amplifier which is a part of the oscillator circuit. The amplified voltage of the mains frequency appears at the output of the stage and beats with the signal developed by the oscillator. This gives what is known as the secondary beats. It is at the frequency of such secondary beats that the output voltage of the instrument pulsates. When this frequency is close to zero, the pointer of the voltmeter or the glowing part of the indicator screen will register a slow oscillation. The compensation, in this case, is obtained by setting the oscillator scale to 50 cps and by adjusting the trimmer C until the secondary beats are slowed down and cease altogether, when F will be equal to 50 cps. The calibration correction should be carried out every time before the instrument is put into operation.

Voltages U_1 and U_2 supplied by the two heterodynes to the detector stage must differ in their amplitudes. The greater the difference of amplitude values of these voltages, the smaller will be the harmonic content in the output voltage U_0, although the output voltage value will also be lower. In practical cases, voltage U_1 is 20-30 times smaller than voltage U_2. For reducing the harmonic content, the harmonics of one of the two heterodynes have also to be kept out of the detector stage. This is accomplished by passing the signal developed by oscillator H_1 through a filter containing tuned circuits adjusted to frequency f_1. The non-linear distortion can be decreased by using push-pull circuits in the detector stage as well as in the stage of low-frequency amplification. The adjustment of the output voltage of the instrument can be conveniently performed as shown in Fig. 271*a*, i.e., by varying the voltage fed to the detector from heterodyne H_1.

In beat-frequency oscillators, measures should be taken to prevent the effect of the so-called "pulling-in", otherwise known as forced synchronisation. This effect takes place as follows. When the difference between frequencies f_1 and f_2 is slight, the oscillations generated by one of the two heterodynes, for instance by heterodyne H_2, can penetrate into the other heterodyne (H_1), making the latter stage generate a signal of frequency f_2, instead of the required frequency f_1. Should this occur, and the frequency of one of the two heterodynes is forced to follow the frequency of the other heterodyne, a zero-beat condition will be

established over the entire bass-tuning range of the beat-frequency oscillator. This will prevent the instrument from developing the required low pitch audio-frequency signals.

The pulling-in effect may be obviated by eliminating all types of coupling between the two heterodynes H_1 and H_2. This is attained by a thorough shielding of the heterodyne stages and by connecting decoupling filters into their supply circuits.

Special care should be taken to prevent coupling between the heterodyne stages through the common detector stage they feed. From this point of view, the push-pull or balanced detector, shown in Fig. 271*b*, offers the best advantage. Here, the two halves of coil L_I and the input capacitances of valves V_1 and V_2 constitute a balanced bridge circuit, the voltages of the two heterodynes

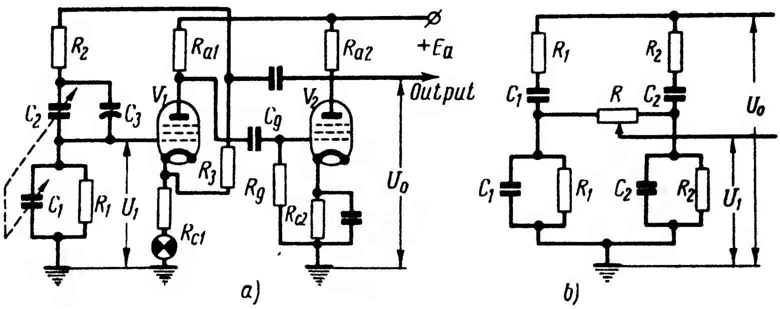


Fig. 272. *RC* audio-frequency oscillator circuits

H_1 and H_2 being applied across the two diagonals of the given bridge. When the instrument has to employ a single-ended detector, a bridge consisting of four similar resistors or capacitors may be provided in the input circuit of such a detector (Fig. 271*c*). Still another way of avoiding this coupling may be provided by the use of a multi-grid valve (for instance, a heptode) in the detector stage. When such a detector is used, the coupling is obviated by feeding voltages f_1 and f_2 to separate control grids of the valve.

It is highly desirable that the inherent stability of both heterodyne stages H_1 and H_2 is made as high as possible. The answer in this case is provided by electron-coupled or transitron oscillators.

RC audio-frequency oscillators. A great popularity is enjoyed by the so-called *RC* audio-frequency oscillators, discussed under this heading. The most widely used circuit diagram of such an oscillator is given in Fig. 272*a*. This is the circuit of a low-frequency two-stage resistance coupled amplifier, in which both the positive and the negative types of feedback are provided between the input and the output. The positive feedback, necessary for the self-excitation of the amplifier, is obtained by means of a special divider comprised of capacitors C_1C_2 and resistors R_1R_2. In one section of this divider, the *RC* elements are connected in parallel with each other (components C_1 and R_1), while in the other section they are connected in series (components C_2 and R_2).

The circuit functions as follows. Assume that a certain voltage pulse U_{g1} reaches the control grid of valve V_1. The phase of this pulse is reversed as the given pulse goes through the process of amplification in the first stage. The inverted phase is then subjected to another inversion as the pulse passes through the second stage. Such double inversion means that the phase of the amplified pulse, appearing at the output, is coincident with the phase the pulse had prior to amplification. In other words, the phase of input voltage U_{g1} is the same as the phase of output voltage U_0. The condition of self-excitation is obtained only when a part of voltage U_0 is fed back to the control grid of valve V_1 in phase

coincidence; in other words, it is required that voltage U_1 of the positive feedback path coincides in phase with voltage pulse U_{g1}. As follows from the oscillator valve theory, such a condition may be obtained only for a certain definite frequency, the frequency depending upon the values of capacitances and resistances used in the positive-feedback divider. Hence, the circuit being described is capable of generating oscillations of only one definite frequency. This is a good thing, because the oscillator of the given type is forced to develop a single sinusoidal frequency, all the harmonic components or other frequencies being kept out of its output. The fundamental frequency of the oscillator may be varied by changing the values of resistances and capacitances of the divider.

In practical cases, capacitors C_1 and C_2, as well as resistors R_1 and R_2, are made equal. Denoting them, respectively, as C and R, we see that the value of R is equal to $\frac{1}{2\pi fC}$, where f is the frequency of oscillations. If follows from this that: $f = \frac{1}{2\pi CR}$.

If the positive feedback divider is comprised of two similar types of resistance (e.g., ohmic resistances), the condition necessary for self-oscillation will exist at any frequency, regardless of the values of the impedances (or resistances) making up the divider. Such a circuit would generate non-sinusoidal voltages, rich in high harmonics. The application of the special divider, however, is still insufficient to obtain pure sinusoidal oscillations. The difficulty is that this divider reduces the voltage by only three times, i.e., U_1 is equal to $\frac{U_0}{3}$. But the total gain of the two stages is much higher than three. Therefore, the control grid of valve V_1 receives a very high voltage, this voltage overloading the grid and causing a strong non-linear distortion. A negative feedback with the factor β only slightly less than one-third is used to limit the amplitude of oscillations to the linear part of the valve characteristic. If $\beta = \frac{1}{3}$, the negative feedback fully compensates for the positive feedback and the oscillation does not take place. If the value of β is smaller, the positive feedback will predominate and, as a result, the stage will oscillate.

The negative feedback decreases non-linear distortion and improves the stability of the oscillator. The negative feedback circuit is constituted by resistors R_3 and Rc_1. The coefficient of the negative feedback is given as:

$$\beta = \frac{R_{c1}}{R_3 + R_{c1}}.$$

Resistor R_3 must be made variable so that the required value of β could be set. A small incandescent lamp is employed as a part of cathode resistor R_{c1} in order to improve the stability of the oscillator and to decrease the non-linear distortion. As the current increases, the resistance of the lamp filament also increases. This property of the lamp is utilised in the following manner. If for one reason or another the amplitude of oscillations grows larger, the current flowing through the lamp will be increased. The higher temperature of the lamp filament, caused by such current increase, will cause an increase of the resistance of the filament. This, in turn, will increase the value of β and, as a result, the amplitude of oscillations will be made smaller. Since the normal operation of of the described circuit requires that the resistance of the lamp is considerable, it is customary to use several small incandescent lamps connected in series. 26-volt 0.15-ampere lamps or else miniature low-power 220-volt lamps are the best ones to employ for the purpose.

The frequency-determining formula given above shows the big advantage the RC oscillator has over the oscillator of the LC type. When it is required to change the frequency from 20 to 20,000 cps, i.e., by 1,000 times, the product LC has to be varied 1,000,000 times in the LC oscillator. On the other hand, when dealing

with the RC oscillator, it is only necessary to vary the RC product 1,000 times to obtain the same frequency change. The latter range of change is much easier to attain in practice. Ganged capacitors C_1C_2 readily secure a 10-fold variation of capacitance. A further extension of the frequency range is provided by resistors R_1 and R_2. A resistor of this type actually consists of three resistors, each resistor differing from the next one by 10 times in resistance value. These resistors are cut in and out of the circuit by the band switch, such switching arrangement offering a simple way of extending the audio-frequency range.

The said frequency-setting resistors are easy to calculate, as shown by the following example. Assume that the capacitance of each variable capacitor in the C_1C_2 gang varies from 40 to 400 pf (these capacitance values include the distributed capacitance of the associated wiring). Keeping in mind that the greatest resistance corresponds to the frequency of 20 cps and to the maximum capacitance of 400, we proceed to find the value of such resistance in the following manner:

$$R = \frac{1}{2\pi fC} = \frac{1}{6.28\times 20\times 400\times 10^{-12}} \approx 20\times 10^{6} \text{ ohms} = 20 \text{ megohms}.$$

When the band switch cuts in this resistor into the circuit, the band covered by the gang capacitor will extend from 20 to 200 cps. The 200-2,000 cps band calls for a 2-megohm resistor, while the 2,000-20,000 cps band is covered when a 0.2-megohm resistor is cut into the circuit. Thus, the entire audio-frequency range (20-20,000 cps) is covered by three bands.

The capacitances of C_1 and C_2 are made equal by adjusting trimmer C_3 connected in parallel with C_2 (Fig. 272*a*). For the sake of simplicity, screen-grid circuits are not shown in the diagram of Fig. 272*a*. In many oscillators of the type here described, the second stage is followed by an additional audio-frequency amplifier stage. The adjustment of the output voltage is usually performed by a potentiometer in the output circuit of the second stage. The oscilator is generally provided with a transformer output. There are, as a rule, two secondary output windings in such a transformer, one winding consisting of many turns and designed to feed high-impedance loads, while the other winding is of a low-impedance type and consists of a small number of turns.

The easy coverage of the entire audio-frequency range is an advantage of the RC oscillator over LC oscillators. Its another advantage is the absence of clumsy iron-core coils in its circuit (such coils are typical of LC oscillators). And, lastly, still another advantage is the better frequency stability, as compared to the frequency stability of beat-frequency oscillators.

The RC oscillator has its own shortcoming, too. This shortcoming is seen in the necessity of using the twin variable gang capacitors in the oscillator circuit. The peculiar way in which these capacitors are connected causes the following inconvenience. If the rotor of the gang capacitors is connected to the grid of valve V_1, the capacitor frame has to be insulated from the oscillator chassis. Moreover, the rotor shaft cannot be brought out to the front panel because of the effect of the so-called "hand capacitance". This undesirable effect is observed when the operator's hand is brought close to the "floating" rotor shaft. The external capacitance of the operator's body introduces an auxiliary capacitance into the oscillatory circuit, thus varying its parameters. This, in turn, varies the frequency generated by the oscillator. The effect of the described hand-capacitance may be eliminated by earthing the rotor of capacitor C_1, but then this rotor will have to be insulated from the rotor of capacitor C_2

Examining various possible circuit arrangements of RC oscillators in general, we note that the circuit shown in Fig. 272*b* offers a better way of arranging the positive feedback than does the circuit of Fig. 272*a*. In the arrangement shown in Fig. 272*b*, two dividers with fixed values of C and R are employed on each frequency band. For instance on 20-200 cps band, one of these two dividers is designed for 20 cps, while the other — for 200 cps. In this circuit, a continuous frequency coverage, within the given band, is obtained by moving the slider

of potentiometer R, the slider being connected to the grid of the valve V_1. When it is required to go over to the next audio-frequency band, it is only necessary to disconnect the end of the potentiometer from the 20-cps divider and to connect it to the following divider, the latter being designed for 2,000 cps. In a circuit of this type, the value of R must be at least seven times as large as the values of R_1 or R_2. Because of this, R_1 and R_2 are made considerably smaller in this circuit, as compared to the previously-studied circuit, while the values of C_1 and C_2 are made respectively greater.

The circuit just described has the disadvantage that its output voltage varies with the position of the slider of potentiometer R. The voltage is at its minimum value when the slider is set to its middle position.

Any one of the discussed audio-frequency oscillators can find a large variety of applications in a radio laboratory. For instance, such oscillators are widely used for testing and adjustment of low-frequency amplifiers. When such an amplifier is tested, the procedure is as follows. First, a voltage divider is connected at the input of the amplifier, the output circuit being loaded with a normal value of resistance. A rectifier-type voltmeter or else a valve voltmeter is used to measure the input and output voltages of the amplifier. When the amplifier is thus set up for the test, the audio-frequency oscillator is switched on. The output of the oscillator is connected to the voltage divider provided at the input of the amplifier, and the oscillator gain is adjusted to feed a small value of signal voltage to the grid circuit of the first amplifier stage. The voltmeter is then brought into play and the values of the amplifier input voltage U_i and output voltage U_0 are measured. Dividing U_0 by U_i gives the overall gain of the amplifier under test. The same oscillator may be also employed for the purpose of determining the frequency characteristic of the amplifier, as well as for the purpose of checking the operation of the gain control and the tone control of the amplifier on various frequencies. If an electron oscillograph (see Sec. 125) is employed with the described testing set-up, it becomes possible, by observing the shape of the output voltage of the amplifier on the screen of the oscillograph, to determine the magnitude of non-linear distortion introduced by the amplifier into its output circuit.

Apart from testing the low-frequency amplifiers, the audio-oscillators find a variety of other applications. They are used to test modulators of radio transmitters, to supply measurement bridges with a.c. voltage, etc.

124. SIGNAL GENERATORS

High-frequency heterodynes, capable of providing a modulated output signal and generally referred to as signal generators, are widely used for testing and alignment of radio receivers. Such signal generators are also employed for many other purposes.

The signal generator simulates the signals of radio stations operating on various frequencies, possessing different output power, and located at different distances from the radio receiver under test. It is generally preferable to test radio receivers with the aid of such signal generators, instead of testing them by tuning to the carrier frequency of an actual radio station. This is appreciated, if we consider that a radio station is not always on the air. Besides, when the station is transmitting speech or music, the amplitude of the received signal is constantly varying. And, finally, the station signal may be fading. All this makes it difficult to perform various measurements in a radio receiver tuned to an actual radio station. Hence the preference given to signal generators. The signal generator develops a signal of the required carrier frequency, and this signal possesses a constant amplitude and modulation percentage.

Fig. 273 illustrates how the signal oscillator should be connected to the receiver under test. This is a block diagram, where the signal oscillator is shown feeding the input circuit of the receiver with a very small value of modulat-

ed high-frequency voltage. The voltage is fed through a dummy aerial, the latter represented by a tuned circuit simulating the receiving aerial. On medium waves, the parameters of the usual receiving aerial are as follows: $C_A = 200$ pf, $L_A = 20$ mh, $R_A = 25$ ohms. Since the inductance of the aerial is of no practical significance, it may be considered to be equal to zero. On short waves, the parameters of the dummy aerial may be taken as: $C_A = 200$ pf and $R_A = 400$ ohms. Measuring electromotive force E_A at the input of the receiver, when such e.m.f. is sufficient to let the receiver develop a normal value of output voltage U_0, determines the sensitivity of the set. When the frequency of the signal generator is varied or else, when the receiver is detuned within small limits, the change of U_0 is a good indication of the selectivity of the receiver.

It is convenient to determine the selectivity of a radio receiver in the following way. After the receiver has been tuned exactly to the frequency of the signal generator, the latter is detuned by, say, 10 kc. This results in a certain decrease of the receiver output voltage U_0. Following this, the output voltage of the detuned signal generator is increased until the former value of U_0 is obtained. If the generator voltage (E_A) had to be increased 20 times to attain this condition, this will indicate that a 10-kc detuning results in a 20-fold signal attenuation. The complete selectivity measurement procedure consists of various degrees of detuning of the signal generator.

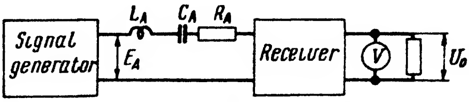


Fig. 273. Testing a radio receiver with the help of a signal generator

The sensitivity and selectivity of a radio receiver is usually checked on at least three points of the receiver tuning range—at the extreme frequencies and in the middle of the tuning scale. During the receiver alignment process, the signal generator is used to adjust the high-frequency and intermediate-frequency circuits, to match the tuned circuits, to check the operation of the automatic gain control circuit and of the manual gain control, etc.

The signal generator must provide a required output voltage over the necessary frequency range and possess a good stability of frequency and calibration. The voltage generated by the instrument must be adjustable between 1 μv and several fractions of a volt, the calibration error of the voltage not exceeding 20%. The signal produced by the instrument should be modulated by sinusoidal 400-cps audio oscillations, the modulation percentage being 30%. Laboratory-type instruments of this kind have a more complex design and are known as *standard signal generators (SSG)*. SSGs are noted for certain additional features which are not possessed by the usual signal generators employed in radio servicing shops. Thus, an SSG is usually provided with some means of varying the modulation percentage.

Fig. 274 gives the block diagram of a typical signal generator. A frequency-calibrated high-frequency oscillator O is the basic part of such a generator. For the sake of better frequency stability, the oscillator employs a two-stage circuit arrangement. Alternatively, an electron-coupled oscillator or a transitron oscillator may be used in this application. Audio oscillator AO, generating a standard average frequency of 400 cps, acts as the modulator. This audio oscillator usually consists of a single inductive-feedback oscillator stage and employs a two-stage circuit arrangement. Alternatively, an electron-coupled oscillator or a transitron oscillator may be used in this application. Audio oscillator AO, generating a standard average frequency of 400 cps, acts as the modulator. This audio oscillator usually consists of a single inductive-feedback oscillator stage and employs a single valve.

Switch S, when thrown to the "off" position, allows the signal generator to develop an unmodulated voltage. This is desirable when testing radio receivers designed for the reception of c.w. signals. Should it become necessary to

modulate the output signal at a frequency other than 400 cps, an external audio-frequency modulating voltage may be applied to terminals "External modulation", shown in the circuit diagram. The modulation percentage may be varied by connecting a potentiometer after switch S. A better stability of the high-frequency is secured when the modulation is applied to the second stage of high-frequency oscillator O. If the oscillator employs an electron-coupled circuit, the modulation is best applied to the suppressor grid of the valve.

High-frequency oscillations developed by oscillator O are fed to potentiometer R. The adjustment of the slider of this potentiometer provides a continuously-variable voltage output of the instrument. The voltage taken off the potentiometer slider is constantly measured by valve voltmeter VV. This voltage is applied to an attenuator consisting of several fixed resistors and step-switch SS.

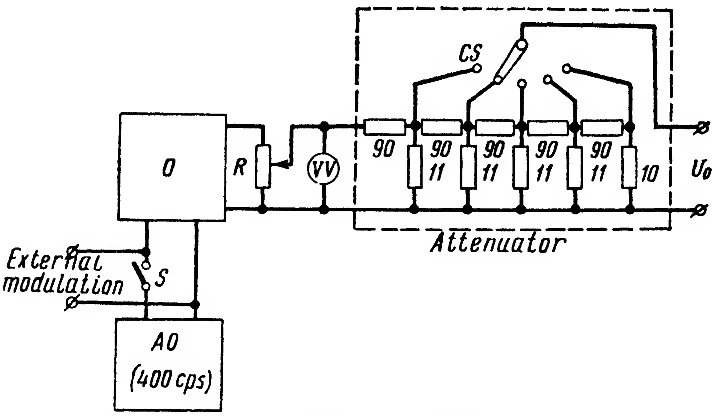


Fig. 274. Block diagram of a signal generator

The circuit diagram gives approximate values of the resistors making up the attenuator. When the moving contact of the step-switch is shifted from one of the stationary contacts to the next stationary contact, the voltage across the output terminals of the instrument is decreased by 10 times. This voltage (U_0) is equal to 0.1 of voltage U read by voltmeter VV when the moving contact is set to the first stationary contact. When the moving contact is set to the next stationary contacts 2, 3, etc., U_0 becomes equal, respectively, to 0.01 U, 0.001 U, etc. If potentiometer R is so adjusted that VV gives a reading of $U =$ $= 1$ v, voltage U_0 across the output terminals of the instrument will be equal to 10 microvolts when the moving contact of switch SS is set to the 5-th stationary contact. The value of U_0 is always determined by multiplying the voltmeter reading by the division factor corresponding to a given position of switch SS.

Proper design of this switch is somewhat of a problem, because the resistors comprising the attenuator must possess the smallest possible values of distributed capacitance and inductance. Such parasitic capacitances and inductances can introduce large measurement errors on short waves. To obviate this, the resistors of the attenuator are wound with thin rheostat wire and are so made that the winding possesses no inductance.

The signal generator can employ a simple diode valve voltmeter, employing a circuit such as that shown in Fig. 261d. The simpler types of such generators are built without a voltmeter of any kind. In view of the voltmeter, potentiometer R used in a generator of such type is calibrated on voltage values, for instance — from 1 to 0.1 v. In such a circuit, when the generator output voltage, fed to the potentiometer, is adjusted to a constant value of 1 volt, the value of U_0 across the output terminals of the instrument may be determined without

a voltmeter by multiplying the voltage division factor of the potentiometer by the division factor of the attenuator. For instance, if the potentiometer knob is set to position 0.35 and the attenuator knob to position 0.0001, the total division factor will be equal to 0.000035 and, hence, $U_0 = 35$ microvolts.

It is important that high-frequency oscillations, developed by the signal generator, reach the radio receiver under test only through the potentiometer and the attenuator. A direct pick up of the generator signal by the radio receiver will be misguiding. In order to preclude the possibility of such a pick up, the whole of the signal generator must be thoroughly shielded. Besides this, appropriate filters, connected into the power supply circuits of the signal generator, must be employed for the purpose of suppression of high-frequency radiation. Unless such measures are taken, the high-frequency oscillations will reach the radio receiver circuits not only through the attenuator, but also through other paths, for instance—through stray capacitive couplings. As a result, the value of high-frequency voltage applied to the input circuit of the radio receiver may be considerably greater than the value computed from the division factors. As a result, the measurements obtained under such conditions will be highly erroneous.

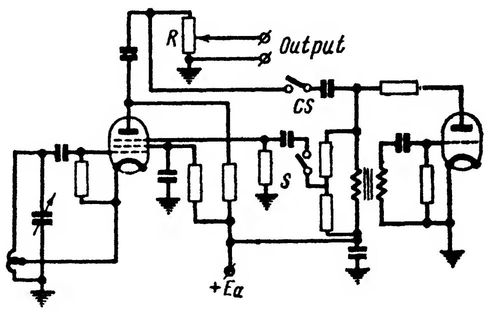


Fig. 275. Circuit diagram of a simple modulated heterodyne

It is a good practice to provide an additional switch in the signal generator, such a switch disconnecting the high-frequency oscillator and connecting audio oscillator *AO* directly to the voltage divider circuit (i.e., to the potentiometer and to the attenuator). This done, the instrument may be used for testing low-frequency amplifiers. The input circuit of such an amplifier will, in this case, be connected directly to the terminals of the instrument, the necessary audio-frequency low potential being developed across the terminals.

It is possible to design a very simple type of signal generator employing no attenuator. Although such instrument cannot be used for the measurement of sensitivity and selectivity of radio receivers, still it can be employed for the alignment and adjustment of the radio sets. Such simplified version of the signal-generator must be also subjected to a thorough shielding, and the instrument should be located at a certain distance from the radio receiver under adjustment. As a rule, the receiver will need no aerial to pick up the signal developed by such a generator, because the signal is radiated by the output terminals of the instrument. Should it happen that the receiver sensitivity is very low and the signal cannot be heard on the radio set, a short piece of wire connected to the output terminal of the generator or to the aerial terminal of the receiver will boost up the signal strength to a sufficiently high degree for the measurement.

When the simplified generator is to be used for testing an intermediate-frequency stage of the radio receiver, the output terminal of the instrument is connected by a piece of wire directly to the control grid of the respective i.f. valve. In order to cover the tuning range of an "all-wave" radio receiver, the instrument should be capable of generating signals in the frequency range of 100 kc-30 mc. Fig. 275 gives the circuit diagram of a simplified modulated-wave signal generator. The h.f. oscillator employs an electron-coupled circuit. The audio-frequency oscillator, employing an inductive-feedback arrangement, develops the modulating signal and applies it to the suppressor grid of the h.f. valve. For the sake of simplicity, the band-switching incorporated in the instru-

ment is not shown in the circuit diagram. In this instrument, the modulation s stopped by switch S, while the change-over switch CS makes it possible to connect the audio voltage to output potentiometer R, when the instrument is used for audio-frequency testing purposes only. In the latter use, the high-frequency oscillator should be switched off.

125. ELECTRON OSCILLOGRAPHS

The electron oscillograph, known in radio-engineering simply as an oscillograph or an oscilloscope, is the most universal testing instrument. The oscillograph offers a possibility of not only making various types of measurements, but also of visual observation of various processes taking place in radio equipment.

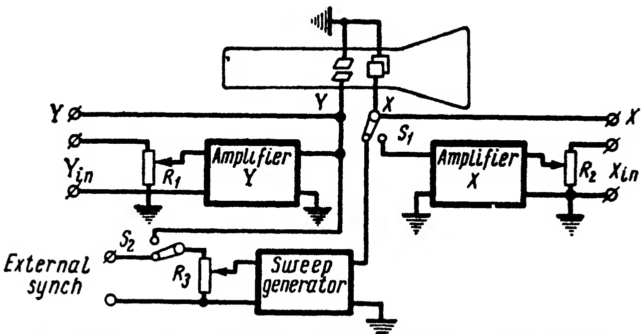


Fig. 276. Block diagram of an electron oscillograph

ment. The screen of the oscillograph gives a direct glowing representation of alternating voltages and shows the action of electron valves, tuned circuits and amplifiers. Such visual representations, appearing on the oscillograph screen are called *oscillograms*.

The block diagram of an oscillograph is given in Fig. 276. To simplify the power supply circuits feeding various electrodes of the cathode-ray tube (CRT) are not shown in the diagram. This is quite justified because we already know the general design and operational features of CRT from Sec. 50.

Referring to the block diagram, we see that the alternating voltage being tested or measured is applied across input terminals Y_{in}. From these terminals, through potentiometer R_1 and amplifier Y, this voltage is applied to plates Y of the CRT. The magnitude of the voltage at Y is adjusted by means of potentiometer R_1. If the voltage being measured is sufficiently high, it can be supplied to the plates of CRT directly through terminal Y, without the preliminary amplification. A circuit arrangement just described is repeated at the right-hand side of the drawing and relates to feeding voltages to plates X of the CRT.

By means of the sweep change-over switch S_1, plates X may be connected to the local sweep-voltage generator. This allows us to observe the curves of alternating voltages on the CRT screen. From plates Y, the voltage being analysed (or some other external voltage, applied to the terminals of external synchronisation) is fed to the sweep-voltage generator through switch S_2. The value of the synchronising voltage is adjusted by means of potentiometer R_3.

The amplifier employed by the oscillograph is usually represented by a single resistance-coupled stage. A choke is sometimes connected in series with load resistance R_a of this stage in order to decrease the attenuation of the frequency characteristic on the higher frequencies. This permits the amplifier to function normally on frequencies as high as 50-100 kc. In order to provide high values

of the input resistance, potentiometers R_1 and R_2 are of the 1-megohm variety. Thanks to this, when the oscillograph is connected to the circuits being analysed, these circuits are not loaded heavily by the potentiometers. The amplifiers increase the sensitivity of the oscillograph several dozens of times. For instance, if the sensitivity of *CRT* is 0.2 mm per volt, it will be increased up to 4 mm per volt when an amplifier with a gain of 20 is incorporated in the circuit. When the oscillograph design must be as simple as possible, both amplifiers or one of them (the one used with X plates) are eliminated from the instrument. If low voltages are to be measured with such a simplified oscillograph, external amplifiers will have to be used with the instrument.

The required sawtooth voltage is fed to plates X from the sweep-voltage generator; the shape of this voltage is shown in Fig. 277*a*. During the span of time, denoted as t_1, when the sweep voltage increases, the electron ray (and hence the light spot on the screen of the *CRT*) moves at a constant speed and in one direction — for instance, from left to right along the horizontal plane. When the sweep voltage drops sharply (see Fig. 277*a*) during this span of time (t_2) the ray and the light spot perform a rapid back swing, known as the *flyback*. The described forward and flyback motions of the electron ray and of the light spot are repeated at the frequency of the sawtooth voltage all the time while the oscillograph is functioning.

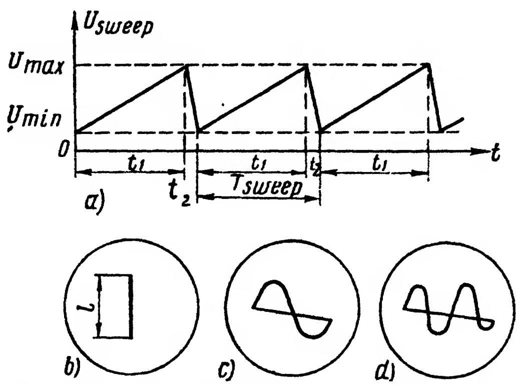


Fig. 277. The shape of sawtooth sweep voltage (*a*) and some of the figures appearing on the oscillograph screen (*b*, *c* and *d*)

If no sweep voltage is present on plates X, the light spot moves only along the vertical plane when an alternating voltage is applied across plates X. In this case, the *CRT* screen displays only a luminous vertical line (Fig. 277*b*). Length l of this line is proportional to double the amplitude of the applied voltage $2U_m$. If the sensitivity of the oscillograph is known, the value of U_m may be determined from the length of the line. For instance, if the sensitivity of the oscillograph is equal to 0.4 mm per volt, while $l = 20$ mm, then $2U_m = \frac{20}{0.4} =$ $= 50$ v; hence, $U_m = 25$ v. Obviously, an oscillograph may be employed as an amplitude voltmeter. This is a very universal voltmeter, because it will measure with equal ease a voltage of any frequency.

When the sweep-frequency voltage is present on plates X and the voltage being analysed is applied across plates Y, the light spot will simultaneously perform an oscillatory motion along the vertical plane and a reoccurring constant-speed forward-and-flyback motion along the horizontal plane. As a result of such complex motion, the screen will display a single luminous curve of the voltage being analysed. A curve of such type is shown in Fig. 277*c*. In this case, *time scanning* is said to take place, — the voltage analysed being scanned.

In order to make the curve remain stationary on the *CRT* screen, period T_s of the sawtooth sweep voltage must be equal to period T of the voltage being analysed or, alternatively, T_s may be a multiple of T. When such conditions are attained, the voltage being analysed will go through a complete oscillation (or through a complete series of oscillations), while the sweep voltage goes through one complete cycle determined by T_s. Under such conditions, at the end of the flyback cycle the light spot will happen to be exactly at the same position from

which it began to move at the moment its forward motion commenced. This results in the stationary position of the curve. Fig. 277c and Fig. 277d give the curves of a sinusoidal voltage for two cases, when $T_s = T$ and $T_s = 2T$. It is desirable that the flyback time t_2 is made as short as possible, because during this time a part of the curve is not reproduced in the screen. Besides this, the shorter the flyback time t_2, the quicker will the electron ray return to its initial position, and the weaker will be the flyback trace displayed on the screen.

The simplest sweep-voltage generator is represented by an oscillator employing a thyratron valve (Fig. 278a). In this circuit, one of capacitors C_1, C_2 or C_3 is charged from a source of d.c. voltage E_a through resistor R. Negatively-biased

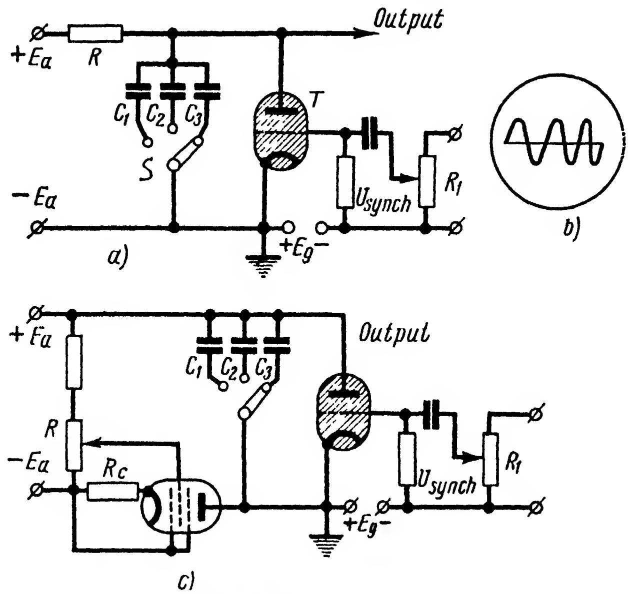


Fig. 278. Sweep-voltage generator circuits with thyratrons

thyratron T is connected in parallel with this capacitor. When the capacitor is being charged, the voltage across it gradually increases. When this voltage reaches the value of U_{max}, equal to the firing-voltage, U_f, of the thyratron, the thyratron will pass current. The internal resistance of the thyratron will fall to a very low value and the capacitor will discharge through it until the voltage across the capacitor will drop to a certain value of U_{min}, at which the thyratron current is cut off. Following this, the gradual charging of the capacitor through the resistor R will again commence, and the process just described will repeat itself, by virtue of which a sawtooth voltage will be built up across the capacitor, which can be applied across plates X. The anode current of a thyratron is cut off when the anode voltage of the valve drops to about 10-20 volts. When a considerable negative voltage is applied to the thyratron grid, firing voltage, U_f of the valve may be as high as 200-300 volts and even higher. These characteristics of the described circuit make it possible to obtain the sawtooth sweep voltage with an amplitude of 200-300 volts, which is quite sufficient to sweep the electron ray right across the *CRT* screen. Frequency *FS* of the sweep voltage is roughly adjusted by means of switch S, this switch cutting in the required capacitor into the circuit. The continuous and finer adjustment of the frequency of this sawtooth voltage is provided by variable resistor R_1. The greater the capacitance of the capacitor and the greater the resistance

value of R_1, the slower will be the charging process and the lower will be the frequency of the sweep. The sweep frequency must be so adjusted that frequency f of the voltage being analysed is a multiple of such sweep frequency.

The circuit shown in Fig. 278*a* is noted for the following disadvantage. The capacitor charges through the resistance unevenly: at first fast, then more slowly. This results in corresponding uneveness of the sawtooth voltage increase and of the movement of the light spot across the *CRT* screen. Because of this, the oscillogram will be distorted and will have an uneven time scale (Fig. 278*b*). The unevenness of the sweep can be reduced by charging the capacitor only up to such voltage value which will be several times smaller than E_a. But then the amplitude of the sawtooth voltage will not be sufficient to sweep the electron ray right across the *CRT* screen. It then becomes necessary to amplify this voltage by means of amplifier *X*.

A better method consists in replacing resistor *R* by a pentode valve, such as 6Ж7, 6K7 or another valve, with which, under normal operating conditions, a change of the anode voltage does not cause any change of the anode current. If the charging current is constant, the voltage across the capacitor will increase proportionally to time, thus providing an even speed of the sweep. At the same time, the method being described secures a sufficient amplitude of the sweep voltage. The circuit diagram of a sweep-voltage generator employing a pentode valve is shown in Fig. 278*c*. In this circuit, a continuous frequency-adjustment is obtained by varying the anode resistance of the valve. This is done by changing the screen-grid voltage of the valve by means of potentiometer *R*. Cathode resistor R_c provides both the initial bias voltage of the pentode and the negative feedback. This feedback secures an additional stability of the anode current.

The frequency of the sweep-voltage generator is not stable and easily changes under the influence of variations in supply voltages, temperature and other factors. When such a change takes place, the multiple relation between frequencies f and f_s will be upset. This will result in shifting and swinging of the oscillogram displayed on the *CRT* screen. Such undesirable effect is eliminated by applying synchronising voltage U_{syn} to the grid of the thyratron valve. When the oscilloscope employs an internal synchronisation, U_{syn} is provided by the voltage being analysed. Because of the presence of the synchronising voltage, the thyratron passes current only during the positive half-waves of the synchronising voltage and only under the condition that voltage U_a across the thyratron is sufficiently high. As a result, the multiple relation between f and f_s is automatically maintained. If f slightly changes for one reason or another, f_s will change correspondingly, but the oscillogram on the *CRT* screen will remain stationary. The action of the synchronisation circuit can be intensified by adjusting R_1, which varies the value of U_{syn}.

It sometimes becomes necessary to synchronise the sweep-voltage generator with the help of an external a.c. voltage. Higher sweep-voltage frequencies are obtained from low-power thyratron filled with inert gases, such as thyratrons type TГ1-0,1/0,3. Such thyratrons provide sweep-voltage frequencies as high as 50-100 kc. When this is insufficient, still higher frequencies may be obtained by the use of electron-valve sweep-voltage generators. The sweep-voltage generator of an oscillograph is usually supplied with power from the same rectifier which feeds the oscillograph amplifiers.

As already stated at the beginning of this paragraph, the oscillograph is a very versatile instrument. We have already seen how it is used for the measurement of voltage amplitudes and for the visual observation of a.c. voltage curves. This, however, is far from all the possible applications of the instrument. The oscillograph can be used for many other purposes, two of which are discussed below.

If, applying the methods described above, we attempt to view a modulated voltage on the *CRT* screen, the high-frequency oscillations will not, as a rule, be seen, because they are merged on the screen and produce a common luminous band of varying width (Fig. 279*a*). The maximum width of the band, denoted by letter *A*, and the minimum width, denoted by letter *B*, correspond,

respectively, to the greatest and the smallest amplitudes of the modulated oscillations. Thus, although we do not see each individual h.f. oscillation on the screen, we can observe the modulation. Moreover, we can determine the modulation percentage by the geometric measurement of the width of sectors A and B and by applying the following formula:

$$m = \frac{A - B}{A + B}$$

where m stands for the modulation factor.

In this measurement, when the frequency of the carrier signal is expressed in hundreds of kilocycles and higher, the signal should be applied directly to

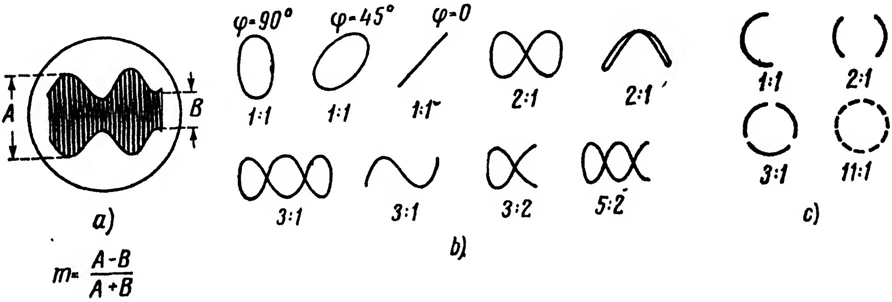


Fig. 279. Modulated oscillations, and figures, displayed on the screen of an oscillograph during comparison of frequencies

plates Y of the *CRT*, because the ordinary amplifier will not pass oscillations of such high frequencies. The stability of the oscillogram of the modulated oscillation can be secured only if the oscillograph is externally synchronised from the modulator.

Another interesting application of the oscillograph is found in the comparison of various frequencies displayed on the *CRT* screen. This method is employed, in particular, in the calibration of audio-frequency oscillators. In this case, an a.c. voltage of some known frequency f_X (for instance 50 cps) is applied to plates X of the *CRT*. This voltage, known as the reference-frequency voltage, can be conveniently obtained from the usual a.c. mains. Plates Y of the *CRT* are fed with the voltage of unknown frequency f_y, generated by an audio-frequency oscillator. The sweep frequency generator is not required in this case.

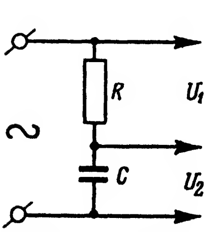


Fig. 280. An *RC* circuit employed to obtain two voltages with a 90° shift

Having prepared the oscillograph circuit in the above manner, we can now vary the frequency of the audio oscillator until the *CRT* screen displays various stationary images, corresponding to one or another relationship of the frequencies. These images are known as Lissajous figures. Some of such figures are shown in Fig. 279*b*.

When the frequencies of the voltages applied to plates X and Y are equal to each other, the *CRT* screen will display a circle, an ellipsis, or a tilted straight line. Which one of the three figures will be displayed, depends upon the phase shift between the two voltages and upon the amplitudes of such voltages. When the frequency relationship is given as 2 to 1, the screen will display a figure resembling number 8 or an arc. This is attributed to the fact that during one oscillation in the horizontal plane two oscillations in the vertical plane will take place.

The application of this method provides a means of measuring several frequencies. However, when the relation $f_y : f_x$ exceeds 10: 1 the figures become too difficult to read. The situation can be amended by employing a higher reference frequency.

An alternative method of frequency comparison on the screen of *CRT* is as follows. Plates *Y* and *X* of the *CRT* are fed with voltages of some known frequency f_s, the two voltages being shifted in phase by 90°. This will cause a circular scanning, i.e., the screen will display a circle when the values of the voltage amplitudes are properly adjusted. If the voltage of the frequency to be measured is applied to the control electrode of the *CRT*, a break will appear in the circle when the frequencies are in the multiple relation (Fig. 279*c*). This is attributed to the fact that the *CRT* will be operating under the condition of cutoff during the negative half-waves of the voltage applied to its control electrode. The number of breaks in the circle will correspond to the relation $f : f_s$.

Fig. 280 gives the circuit arrangement of an *RC* net by means of which two voltages U_1 and U_2, shifted in phase by 90°, may be obtained.

126. FREQUENCY MEASUREMENTS

In radio engineering practice, an accurate measurement of frequency is a matter of paramount importance. Frequency (or wavelength) measurements are a necessity in the adjustment and maintenance of high frequency oscillators and radio transmitters.

An equally important role is attached to frequency measurements when it is desired to calibrate a radio transmitter, receiver or an oscillator — and to make sure that the calibration stays accurate.

There are many methods of performing frequency measurements and many specially-designed instruments to cope with the task. Perhaps the most popular among such instruments are the ones intended for the direct measurement of wave-length or frequencies. These devices are known as the wavemeters, whose basic circuits are discussed in this article.

Resonance wavemeters. Several circuit versions of resonance wavemeters are shown in Fig. 281. Basically, the resonance wavemeter is nothing but a tuned circuit provided with some type of resonance-indicating device.

The tuned circuit of the wavemeter must be designed to provide a high frequency stability. Because of this requirement, the coil and the capacitor used by the wavemeter are made mechanically strong to ensure that they retain their rated inductance and capacitance values. In order to obtain a sharp resonance, the electric qualities of the tuned circuit must be high, too. If the wavemeter is designed for operation over a wide frequency range, it is sometimes supplied with detachable coils. When the wavemeter is calibrated in frequency or wavelength, the calibration is usually entered into a table or a chart although occasionally it is recorded right on the instrument scale.

When the wavemeter is to measure the frequency of a radio transmitter or an oscillator, the instrument must be coupled to the high-frequency generating circuit in one way or another. The most common way of coupling is that of bringing the wavemeter coil in the proximity of the aerial or of the tuned-circuit coil of the transmitter, taking care that the coupling is not made too close, otherwise excessive current might flow through the wavemeter indicator and damage it. When dealing with a comparatively high-power radio station, it is often sufficient to set down the instrument side by side with the transmitter in order to obtain a reading on the wavemeter indicator. In order to secure an accurate frequency reading, the variable capacitor of the wavemeter tuned circuit is rotated until the indicator gives a maximum reading. This is the condition of resonance, and a look at the wavemeter dial with a subsequent reference to the calibration chart or table will ascertain the exact frequency at which the

radio transmitter or oscillator is operating. It sometimes becomes necessary to adjust a radio transmitter to an assigned wavelength, the wavemeter serving as a reference. In such a case, the wavemeter dial is set to the required frequency and the frequency of the transmitter is varied until the wavemeter gives a maximum reading.

Wavemeters can employ various types of indicators. The most desirable type of such indicator is some type of a meter, such as a thermocouple galvanometer, or else a galvanometer supplied with a semiconductor rectifier or a diode valve. Indicators of this type are shown in Fig. 281 *a*, *b* and *c*.

Where simplicity is the object, the meter-type indicator may be replaced by a miniature incandescent lamp or by a neon lamp (Fig. 281 *d* and *e*). However, when the wavemeter is to show the condition of resonance by the

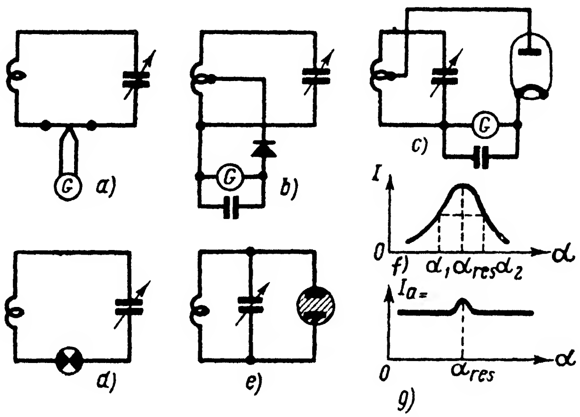


Fig. 281. Circuit diagrams of resonance-wavemeters with different types of indicators, and graphic representation of resonant condition

maximum value of the current flowing through the indicator, the glowing types of indicators cannot provide a very high accuracy of reading. When a meter-type indicator is used, the highest accuracy is obtained by the *double-reading method*, consisting in the following.

The wavemeter dial is lightly detuned both sides of the exact resonance position, the indicator giving similar readings on both points α_1 and α_2. The described procedure completed, two values (α_1 and α_2) will be obtained ons the calibration chart of the wavemeter, as shown in Fig. 281*f*. Referring now to the two recorded readings α_1 and α_2, the actual resonance reading α_{res} may be computed with great accuracy as the mean value of α_1 and α_2, i.e.:

$$\alpha_{res} = \frac{\alpha_1 + \alpha_2}{2} .$$

When a wavemeter is used to measure the frequency of a low-power oscillator the wavemeter indicator may not be sufficiently sensitive to give a reading. If such be the case, the condition of resonance will have to be determined by the so-called *reaction method*, the nature of which is as follows. When the tuned circuit of the wavemeter is coupled to the tuned circuit of the oscillator, whose frequency is to be measured, the moment when the two circuits are brought into resonance is indicated by a change of d.c. component $I_{a=}$ of the anode current or by a change of grid curent $I_{g=}$ of the oscillator. The milliammeter connected into the anode circuit of the stage gives the most noticeable deflection in this resonance-indicating method. During this measurement, the anode

current of most oscillators slightly increases during the moment of resonance, as shown in Fig. 281*g*.

The resonance wavemeters are noted for the following serious shortcomings. They are not suitable either for the calibration of radio-receiver scales, or for the accurate measurement of the frequency to which the receivers are tuned. The only exception, in this case, is the regenerative receiver, which, under regenerative condition, can be calibrated with the help of a resonance wavemeter, and whose tuning frequency may also be determined by the same instrument. The accuracy of measurements is not high with wavemeters; even the best of them give an error as high as 0.1% in measurement, while the error of the simpler resonance wavemeters can be 0.5% and even higher. All this is explained by the comparatively close coupling that has to be used between the wavemeter and the circuit under test. This changes the parameters of the tuned circuit of the wavemeter, and the closer is the coupling, the flatter becomes the resonance curve of the instrument. Converselv, when the wavemeter is used to measure the frequency of a self-excited oscillator, the proximity of the tuned circuit of the wavemeter to the tuned circuit of the oscillator will change the oscillator frequency. Many modern oscillators and transmitters possess a higher frequency stability and a more exact calibration than the resonance wavemeters. Hence, there is no sense in measuring the frequency of such oscillators and transmitters with the aid of the usual resonance wavemeters and the latter have been largely replaced by the more precise heterodyne wavemeters, described below. The only domain where the resonance wavemeters are still widely used is the range of extremely high frequencies (centimetric and decimetric waves).

Heterodyne wavemeters and crystal calibrators. The heterodyne wavemeter is, in effect, a low-power valve oscillator designed for operation over a continuous frequency range and calibrated in frequency or wavelength (the calibration may be entered in charts or tables, or else, may be provided directly on the instrument scale). The frequency stability of the instrument must be very high. The wavemeter usually employs some variety of the electron-coupled oscillator circuit. The electron-coupled transitron circuit is frequently used in this application; and the instrument often uses two stages (see the transitron oscillator circuit in Fig. 183). The tuned circuit of the heterodyne wavemeter must have very stable parameters and be of high quality. And it is desirable to have very stable power supply.

The heterodyne wavemeter is suitable for frequency measurements in radio receivers and offers a much higher measurement accuracy than that provided by the resonance wavemeters. Such high accuracy of the instrument is attributed to the fact that the heterodyne wavemeter is used to determine the state of resonance by the zero-beat method, this method offering a very high precision of measurements, even when the wavemeter is very loosely coupled to the circuit under test—be it the circuit of a radio transmitter or radio receiver.

When the calibration of a receiver is carried out with the aid of the heterodyne wavemeter, or when such calibration is checked with the help of the same instrument, the latter is set up side by side with the receiver. The radio receiver should be disconnected from its aerial and must operate under the condition of c.w. signal reception, i.e., when the receiver feedback is brought up to the point of oscillation, or else — when the receiver is a superheterodyne set — when the beat-frequency of the receiver is switched on. The calibration of the receiver is carried out in the following manner.

First, the wavemeter dial is set to the required frequency, the wavemeter radiating a weak signal at the given frequency. After this, the receiver tuning is varied until the set picks up the signal generated by the wavemeter, as indicated by the appearance of a beat note in the earphones or in the loudspeaker connected to the receiver. Now, the receiver tuning dial is rotated very slowly until the beat note disappears, after its lowest pitch has been reached and passed. This is the zero-beat condition, indicating that the receiver is tuned to the exact frequency of the heterodyne wavemeter.

The slight measurement error that can be experienced during this type of measurement is attributable to the fact that the human ear cannot perceive the sounds of very low frequency — a few cps — near the point of the precise zero-beat condition. Another reason for the error is the pulling-in effect that can be occasionally experienced in some types of heterodyne wavemeter circuits. However, when the coupling between the measuring instrument and the radio receiver is made very loose (which is always the case), the said measurement error does not exceed a few dozens of cycles per second, which is of no practical significance.

When the described wavemeter is used to measure the frequency to which the radio receiver is tuned, the procedure is the same as explained above, except that the receiver tuning is left untouched, while the wavemeter tuning is varied to obtain the zero-beat condition. If the radio receiver has no provision for the reception of c.w. signals (i.e. when the set does not employ regeneration or a beat-frequency oscillator), the zero-beat method is not applicable. (This pertains to practically all home radio receivers, designed, as a rule, for the reception of modulated signals only. — *Translator's note.)* With such receivers, it becomes necessary to modulate the carrier generated by the heterodyne wavemeter. The carrier should be modulated by a frequency within the range of 400-1,000 cps, the source of modulating voltage being an ordinary audio-frequency oscillator. When no such oscillator is available, the carrier can be modulated by the 50-cps frequency obtained from the power mains. In this method, the state of resonance between the wavemeter and the radio receiver is indicated by the maximum loudness of the modulated signal reproduced by the earphones or by the loudspeaker connected to the output of the radio receiver.

The circuit of the heterodyne wavemeter operates under a class C oscillator condition. Because of this, numerous harmonics are present in the output of the wavemeter. These harmonics may be utilised to expand the operating frequency range of the instrument. This, in its turn, will expand the frequency range within which measurements can be made, whether dealing with receivers or transmitters. Taking the case of a radio receiver, which is to be calibrated on frequencies between 5,000 and 10,000 kc, assume that we have to resort to the aid of a heterodyne wavemeter, whose fundamental frequency range extends only from 150 to 1,500 kc. In such a case, the harmonics generated by the wavemeter have to be used in order to calibrate the receiver over its entire frequency range of 5,000-10,000 kc.

The calibration process under the stipulated conditions is carried out in the following way.

Setting the wavemeter tuning to a frequency of 1,500 kc, we make the instrument generate the following harmonics:

2nd harmonic—3,000 kc,
3rd harmonic—4,500 kc,
4th harmonic—6,000 kc,
5th harmonic—7,500 kc,
6th harmonic—9,000 kc, etc.

When the radio receiver being calibrated is tuned to frequencies of 6,000; 7,500; and 9,000 kc, we obtain zero beats. This makes it possible to perform the receiver calibration at the given frequencies.

The calibration of the receiver may be carried out on the remaining frequencies of its tuning range by setting the heterodyne wavemeter fundamental frequency to, say, 1,000 kc. The instrument will now generate appropriate harmonics to obtain zero beats at the frequencies of 5,000; 6,000; 7,000; 8,000; 9,000 and 10,000 kc (these frequencies correspond, respectively, to the 5th, 6th, 7th, 8th, 9th and 10th harmonics of the wavemeter). Such an abundance of harmonics makes it easy to calibrate the receiver over its entire tuning range.

The reverse problem is just as easily solved, i.e. — the problem of checking the accuracy of tuning of a receiver whose scale had been previously calibrated. Assume that it is required to check the receiver scale at a frequency of 6,500 kc.

Dividing this figure by harmonic numbers 2, 3, 4, etc., we find that when 6,500 is divided by 5 the answer will be 1,300. This is the number of kilocycles to which the wavemeter tuning must be set. This done, zero beats may be obtained in the radio receiver when the set picks up the 5th harmonic of the wavemeter. Assume that, as a result of calibration inaccuracy, the beats are obtained when the wavemeter frequency is equal to 1,310 kc. Consequently, the actual tuning frequency of the radio receiver is not 6,500 kc but rather $1{,}310 \times 5 = 6{,}550$ kc.

The presence of harmonics at the output of the heterodyne wavemeter is a great convenience, because it provides a means of measuring the frequencies of waves shorter than the fundamental waves generated by the instrument. However, great care should be taken to select the right harmonics for the measurement purposes. Should one of the harmonics be mistaken for another, and the receiver is calibrated against the wrongly-selected harmonic, a great calibration error will result. Because of this, a radio receiver should be calibrated against the wavemeter harmonics only when the tuning range of the receiver is approximately known. Only in such a case the calibration process may be begun and carried out in the following way.

First, the wavemeter is set to operate at its highest frequency and the so-called *reference points* are determined. In this stage, the harmonics are distributed in frequency sufficiently far from one another and the danger of mistaking one harmonic for another is non-existent. For further calibration, the wavemeter is tuned to lower frequencies between the reference points. This serves to introduce a larger number of harmonics into the tuning range of the radio receiver.

The higher harmonics are reproduced weaker than the lower harmonics and their level must be raised by connecting a short piece of wire to the receiver or to the wavemeter, the wire serving as an aerial.

The results of calibration (or the results of checking calibration) are entered in a table, an example of which is given below:

Frequency (kc), as read on the wavemeter scale	5,000	4,800	4,700	4,600	4,500
Receiver scale setting	4	8	10	13	16

When this table is completed, a calibration curve facilitating the location of intermediate points is plotted from the table. An example of such calibration curve is given in Fig. 282. A reference to this curve quickly determines a wrongly-plotted point (obtained as a result of confusing the number of a harmonic), because such a point is always set aside of the smooth curve passing through all the other points.

If it is required to measure the frequency or to check the calibration of a radio transmitter, the heterodyne wavemeter is made to operate as a radio receiver. For this purpose, the wavemeter employs a detector stage, which is sometimes followed by a low-frequency amplifier. In simpler versions of heterodyne wavemeters, the heterodyne stage itself is employed as a single-valve regenerative receiver operating under the condition when the threshold of oscillation has just

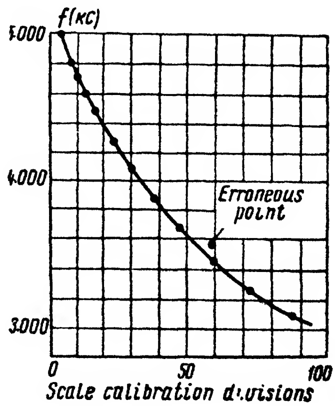


Fig. 282. An example of the calibration curve of a radio receiver

been passed. The earphones are connected right into the anode circuit of the detector stage of the wavemeter. The amplification of the wavemeter operating as a radio receiver is rather low and it, therefore, becomes necessary to connect a short piece of wire to the communication terminal of the wavemeter and to bring the instrument into proximity of the transmitter. In this measurement, the transmitter signal, picked up by the wavemeter, and the signal developed by the wavemeter itself are made to beat in the instrument. The frequency of the transmitter is read off the instrument scale when the condition of zero beats is obtained.

When the harmonics of the transmitter and wavemeter are utilised, it becomes possible to measure the frequencies of transmitter carriers both above and below the fundamental tuning range of the wavemeter. For instance, if the tuning range of the wavemeter extends from 150 to 1,500 kc, while the transmitter has to be tuned to 5,000 kc, the wavemeter scale is set to 1,000 kc. This will result in beating of the transmitter carrier with the 5th harmonic of the wavemeter. Conversely, if a certain high-frequency oscillator is to be tuned to a frequency of 100 kc, the wavemeter scale is set to 200-kc. In this case, the second harmonic of the oscillator, whose frequency is being measured, will beat with the fundamental frequency of the wavemeter. Here, too, the zero-beat condition will be an indication of resonance.

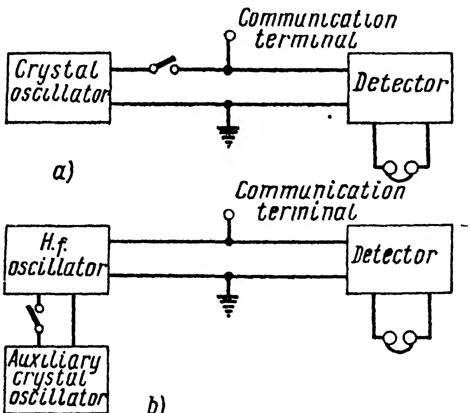


Fig. 283. Block diagrams of a crystal calibrator *(a)* and of a crystal-controlled heterodyne wavemeter *(b)*

It is also possible to beat the harmonics of the transmitter frequency with the harmonics of the frequency of the wavemeter. If the transmitter operates at a frequency of, say, 1,000 kc, setting the wavemeter to approximately the same frequency will cause the strongest beat-frequency note. However, weaker beat notes will be heard at many other settings of the wavemeter dial. Thus, if the instrument scale is set to 1,500 kc, its second harmonic will coincide with the third harmonic of the transmitter; if the wavemeter frequency is equal to 1,250 kc, the fourth harmonic of the instrument will coincide with the fifth harmonic of the transmitter, etc. There are very many combinations of this kind, and this increases the possibility of making mistakes in the determination of harmonic numbers. It should be noted, however, that the higher the harmonic, the weaker will be the resulting audio-frequency beat-note. The following practice is, therefore, recommended, if the mistakes named above are to be avoided. First, the reference points must be found, keeping the coupling of the wavemeter with the transmitter as loose as possible. This will prevent the earphones from reproducing the beats that would otherwise result from the higher but weaker harmonics. Only when this is accomplished, the measuring procedure can be begun without the risk of running into a wrong harmonic.

Besides the heterodyne wavemeters, the so-called crystal calibrators have also found a wide range of application in the measurement practice. Such crystal calibrators resemble the heterodyne wavemeter. The difference between the two types of measuring instruments is seen in that the crystal calibrator employs a crystal control in conjuction with a grid detector (Fig. 283*a*). Because of the fixed tuning, determined by the frequency of the crystal, the crystal calibrator has no continuously-variable frequency range and is used for calibration purposes (or calibration checking purposes) only on some definite frequencies. The

instrument can be employed to calibrate radio receivers, as well as radio transmitters, and to check the accuracy of their calibration.

The oscillator of the crystal calibrator usually employs one of the crystal-control circuits shown in Fig. 186. In this application, the quartz plate is cut in such a way that it has two resonant frequencies, corresponding to the oscillations along the thickness and along the length of the crystal. These frequencies are usually 1,250 and 125 kc, respectively. When the instrument tuning is to be shifted from one of these frequencies to the other, the tuned anode circuit of the oscillator is simply adjusted to either 1,250, or 125 kc, as required.

When the crystal calibrator is used to calibrate a receiver or a transmitter, harmonics developed by the instrument are utilised for the purpose in the same way as they were utilised in the case of the heterodyne wavemeter. The receiver or the transmitter is tuned until the beat-frequency condition is obtained with some harmonic of the crystal calibrator. In order to avoid a mistake in the determination of the number of a harmonic, the reference points are first found by making the crystal calibrator operate at the frequency of 1,250 kc. During calibration of a radio receiver with the help of the crystal calibrator, it is recommended to switch off the calibrator momentarily when the beat-frequency condition is obtained. The disappearance of the sound in the earphones will, then, indicate that the receiver is actually tuned to the calibrator signal, and not to the signal of some radio station.

Because of its high inherent frequency stability, the crystal calibrator provides a better measurement accuracy than does the heterodyne wavemeter. To ensure the highest possible accuracy of measurement, an auxiliary crystal oscillator is sometimes built-in into the heterodyne wavemeter (Fig. 283*b*). Prior to the commencement of a measurement, this auxiliary crystal oscillator is switched on to check the calibration of the heterodyne wavemeter itself. Should it happen that this calibration is slightly off, the required correction can be made by a small trimming capacitor connected into the heterodyne circuit.

If so desired, the usual signal generator can be operated as a heterodyne wavemeter. To do this, it is merely necessary to add a detector stage to the signal generator, or else, to make provision for connecting a pair of earphones into its anode circuit. The output voltage divider becomes unnecessary in this case. Modulation is switched on when radio receivers designed for the reception of modulated signals only are calibrated.

Speaking generally, it is convenient to combine a signal generator with a heterodyne wavemeter, when the extreme simplicity and compactness of measuring instruments is the object. This is quite feasible, because both instruments have a similar basic part—a calibrated heterodyne with a continuously adjustable tuning range.

127. CAPACITANCE AND INDUCTANCE MEASUREMENTS

The inductance of a coil or the capacitance of a capacitor can be measured with the help of a specially devised alternating-current circuit. It is often convenient to employ the standard 50-cps industrial-frequency current in such a circuit, although currents of other frequencies must sometimes be used.

Determining the value of x_C or x_L is the simplest way of measuring capacitance or inductance. This method, however, is not always convenient to use. Whenever possible, x_L and x_C may be measured by the voltmeter-ammeter method or by the method of comparison. Both of these methods were previously explained in Fig. 266 *a*, *b* and *d*, where they were employed in resistance measurements. In the case of a capacitor or a coil, both methods are equally applicable; it is only necessary to feed the circuit with alternating current and to replace resistor R_x by the capacitor or the coil. The meter readings will, then, provide the necessary data for the calculation of x_L or x_C, from which the capacitance or inductance value will be derived.

When choosing between these two measurement methods, we find that the comparison method is more convenient than the other, because it requires but one meter — a voltmeter. The input impedance of such a voltmeter must be considerably higher than either the capacitive reactance x_C or the inductive reactance x_L being measured. A valve voltmeter is the most suitable type of instrument for this measurement.

The measurement is carried out in the following way. Assume that we wish to know the inductance of an iron-core choke. Connecting the choke into the circuit, we apply the comparison measurement method and find that x_L of the choke is equal to 10,000 ohms on the frequency of 50 cps. Resorting to formula $x_L = 6.28fL$, we can now determine the inductance value being sought:

$$L = \frac{x_L}{6.28f} = \frac{10{,}000}{6.28 \times 50} \approx 32 \text{ h.}$$

Thus, we have found the value of inductance. However, it should be noted that the method we have used is applicable only when ohmic resistance r of the choke coil is at least 5 times as small as its inductive reactance x_L, i.e., only when impedance Z of the coil mainly consists of the inductive reactance.

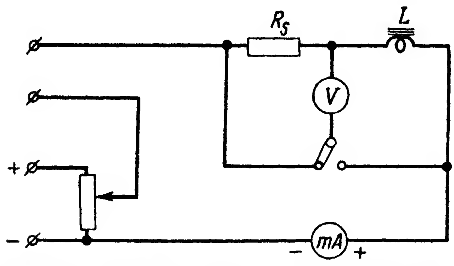


Fig. 284. Measuring the inductance of an iron-core coil

The inductance of an iron-core coil greatly depends upon the value of the magnetising current flowing through it and also depends upon the width of the air gap in the core. In accordance with this, when the inductance value of an iron-core coil is to be determined, it is necessary to employ not only an alternating current for the measurement, but also to pass simultaneously a direct current through the coil. The direct current may be obtained from a rectifier, and the current value must be adjusted (referring to a moving coil milliammeter) to correspond to the proper value of magnetisation for the given type of coil (Fig. 284). In this case, the type of voltmeter employed must possess a closed-circuit input so that the instrument would measure a.c. voltage only.

In preference to the methods described above, the bridge measurement method is more frequently used for the determination of inductance and capacitance values. The operating principle of the bridge circuits employed in this method is similar to the principle of bridge-circuit operation on direct current. Two conditions must be fulfilled to obtain the balance of a bridge circuit on alternating current (Fig. 285a).

The first condition is that the products of impedances of the opposite bridge arms are equal, i.e.:

$$Z_1 Z_3 = Z_2 Z_4$$

The second condition is that the sums of phase-shift angles of the opposite bridge arms are also equal:

$$\varphi_1 + \varphi_3 = \varphi_2 + \varphi_4.$$

The first one of the two conditions is similar to the requirement encountered in any d.c. bridge circuit; when this condition is satisfied in an a.c. bridge, the voltages built up across Z_1 and Z_4 and also across Z_2 and Z_3 will be equal. This, however, is insufficient to obtain the condition of balance in an a.c. bridge and must be supplemented by a correct phase relationship. If no such relationship exists in the bridge, a certain value of a.c. voltage will appear between points A and B. To prevent this, and to obtain the true balance of the a.c. bridge, a phase coincidence across Z_1 and Z_4 and also across Z_2 and Z_3 must also be se-

cured. This is absolutely necessary although meeting the two stipulated conditions often complicates the balancing procedure.

An a.c. bridge may be fed with power from a buzzer circuit. But it is better to feed the bridge from an audio oscillator at a frequency within the range of 400-1,000 cps. Alternatively, the bridge can be powered from 50-cps mains. In the latter case, a step-down transformer must be employed. An earphone is the simplest type of indicator that may be used in an a.c. bridge (Fig. 285).

The sensitivity of such an indicator is sufficiently high. However, when the bridge is supplied with power from the mains, it would be a poor practice to use an earphones as an indicator, because the human ear itself is not sensitive to a frequency of 50 cps.

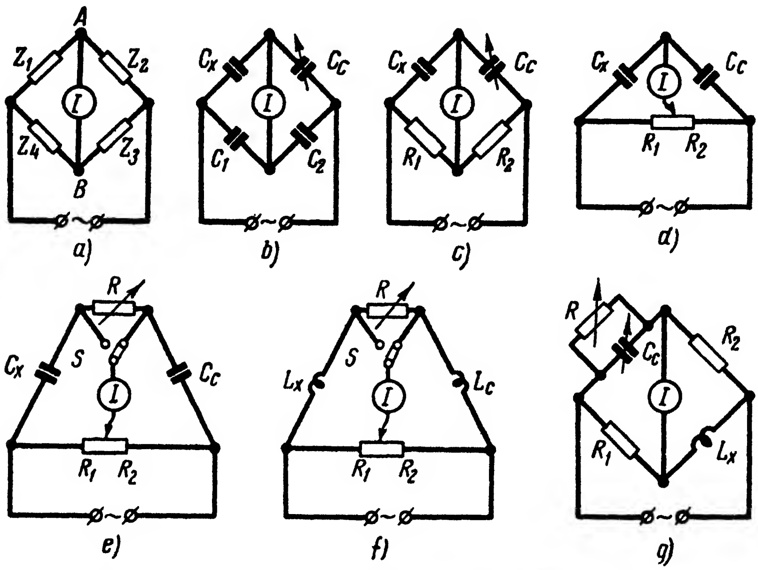


Fig. 285. A.c. bridge circuits for capacitance and inductance measurements

Complex types of a.c. bridges use valve voltmeters or rectifier-type voltmeters as indicators. There is no need to calibrate the scales of these instruments in such an application, because the instruments are used here only to show the minimum value of voltage appearing between points A and B. As a matter of fact, the usual "magic eye", which we know well from our discussion of radio receivers, could be substituted for the voltmeter and would give a good account of itself as a balance indicator of the a.c. bridge circuit. The sensitivity of any indicator employed by an a.c. bridge can be increased by connecting an amplifier between the indicator and points A and B of the circuit.

When balancing an a.c. bridge, the operator should work for the minimum indication (the weakest sound in the earphone or the lowest reading of the voltmeter). It is usually impossible to attain a complete absence of current, flowing through the diagonal-arm, AB, of such a bridge. This effect is attributed to the difficulties encountered in trying to meet the two stipulated conditions necessary for the perfect balance of an a.c. bridge. Another reason why it is practically impossible to get rid of the current flowing through the diagonal, AB, is the inevitable presence of various parasitic capacitances, through which the a.c. voltage manages to reach the indicator. If the supply voltage is non-sinusoidal, its harmonics flow through the parasitic capacitances with particular case.

In this case, the residual current flowing through the bridge diagonal may be quite high, and the dip of the meter pointer (or the minima of sound level in the earphone) will not be sufficiently noticeable during the condition of balance. As a result, the measurement accuracy will be lowered. Besides all this, the conditions necessary for good balance of an a.c. bridge are usually not attained simultaneously both for the fundamental frequency of the voltage being measured and for the higher harmonics of such frequency. As a result, an a.c. bridge is not quite as sensitive nor as accurate as the bridges which we have already studied and which were intended for measurements on direct current only.

Typical circuits of a.c. bridge, designed for capacitance and inductance measurements, are shown in Fig. 285. In circuits given in Fig. 285 *b* and *c*, calibrated capacitor C_c and two standard capacitors C_1 and C_2 (or resistors R_1 and R_2) are employed. Referring to the first one of the two above-stipulated conditions necessary for obtaining the balance of an a.c. bridge, we can write the following formula for the circuit of Fig. 285*b*:

$$C_x = C_c \frac{C_1}{C_2}.$$

In case of circuit of Fig. 285*c*, the formula becomes:

$$C_x = C_c \frac{R_2}{R_1}.$$

When measuring a wide range of capacitances, the bridges of this type are provided with several removable capacitors C_1 and C_2 or with several removable resistors R_1 and R_2. These capacitors or resistors are plugged into appropriate sockets of the instrument. Alternatively, they may be installed inside the instrument-case and connected at will into the circuit by means of a rotary switch. In either case, the resistors or the capacitors are switched into the bridge circuit in such a way, that the ratio $C_1 : C_2$ or $R_1 : R_2$ is equal to $1:1$; $1:10$; $1:100$; etc. For best accuracy of measurements, the ratio is so selected that the condition of bridge balance is secured when variable calibrated capacitor C_c is not set to its extreme positions.

In the circuit of Fig. 285*d*, C_c is a fixed capacitor, known as the standard capacitor because of its exactly known capacitance value. In this circuit, resistors R_1 and R_2 represent a potentiometer, calibrated in terms of the ratio $R_2 : R_1$. The condition of balance of this bridge is similar to the condition of balance of the previously described bridge. In this case, the balance is obtained by a smooth change in the ratio, $R_2 : R_1$, a switch cutting in different fixed-capacitors, C_c, into the circuit, as required by different values of unknown capacitance C_x. The greatest accuracy of measurement is secured when C_x and C_c have approximately the same value, i.e., when ratio $R_2 : R_1$ is close to unity.

Complex types of a.c. bridges always employ an auxiliary variable resistor, which is used for balancing the phase angles. For instance, if capacitor C_x is of a poorer quality than capacitor C_c, the phase-shift angle will be smaller in C_c than in C_x. It is for the purpose of compensating such phase-angle difference that a variable resistor is connected in series with C_c and adjusted until the second condition of balance is obtained. The circuit of such a bridge is shown in Fig. 285*e*. Switch S in the circuit offers a provision for switching in phase-correcting resistor R in series with unknown capacitor C_x, if it happens that this capacitor is of a higher quality than capacitor C_c. When adjusting such a bridge, first an approximate position of balance is found by switching in various values of C_c into the circuit. This done, R is alternately switched between C_x and C_c arms and is continuously adjusted until a sharper point of balance is located. During this procedure, it also becomes necessary to slightly change the value of C_c, which should be followed by a readjustment of R. As already noted above, the described manipulation is rather complex. Still, a careful adjustment of C_c and R should in the end create a condition when the indicator will register a sharp minimum, corresponding to a state of the bridge balance.

We have just discussed a complex a.c. bridge with the resistance compensation of phase-shift losses. It should be noted, however, that in modern capacitors such losses are very low. As a result, even the simplest a.c. bridges, employing no phase-angle compensation, are also capable of sufficient accuracy when used for capacitance measurements. The phase-angle compensation is considerably more important when an a.c. bridge is used for inductance measurements, because inductance coils always possess noticeable ohmic resistance. Fig. 285 also shows bridge-circuit arrangements for the measurement of inductances, the circuit of 285*f* being similar to the circuit of Fig. 285*e*. In this case, the value of L_x is found from the following formula:

$$L_x = L_c \frac{R_1}{R_2}.$$

In order to obviate the necessity of employing standard inductances, the bridge circuit is often arranged as shown in Fig. 285*g*. A circuit of this type can employ the same standards which were used for capacitance measurements. The value of L_x is determined from the following formula:

$$L_x = C_c R_1 R_2.$$

If C_c is a variable capacitor, then R_1 and R_2 are represented by a number of standard resistors. These resistors are cut into the circuit by means of a switch, in accordance with various measurement ranges. However, if C_c is represented by a series of capacitors cut into the circuit at will by means of a switch, then one of the resistors R_1 and R_2 should be capable of continuous adjustment. The equalisation of phase angles is performed by variable resistor R. This component has a high resistance value and is connected in parallel with C_c.

There are many varieties of measuring-bridge circuits. The greatest operational convenience is offered by the so-called universal bridges. These bridges are designed to measure L, C and R and are provided with change-over switches by means of which the most suitable bridge circuit can be employed at will. A.c. bridges possess a rather low measuring accuracy. The error of measurement in such bridges can sometimes be as great as 5 and even 10 per cent, particularly when the bridge is called upon to measure low capacitances and inductances.

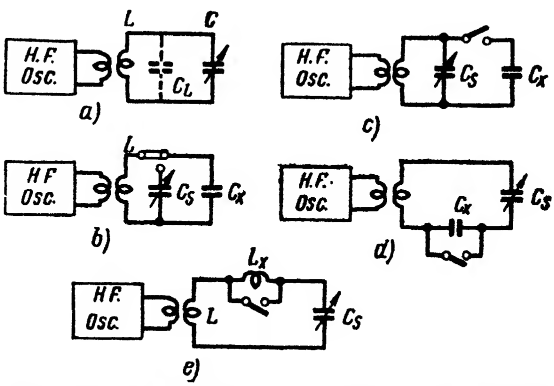


Fig. 286. Various circuits employed in capacitance and inductance measurements on high frequencies by means of resonance methods

It is best to measure such low capacitances and inductances by the methods of resonance. Components of this type are generally used in high-frequency tuned circuits. Hence, it is only normal that a low-capacitance capacitor or a low-inductance coil is connected into a tuned circuit for the purpose of measurement. If the remaining components employed by the circuit have known standard values, and if the circuit is energised from a high-frequency oscillator, the value of the unknown capacitor or coil can be determined by obtaining resonance, and reading off the frequency value from the signal-generator dial. The unknown capacitance or inductance value can then be calculated.

A circuit, in which the value of C_x is measured against standard inductance L_s, or, conversely, L_x is measured against C_s, is given in Fig. 286*a*. Varying

the frequency of the oscillator, or else changing the capacitance, a state of resonance is obtained, whereupon capacitance C_x is calculated from the value of standard inductance L_s, or vice versa. The following set of formulas are applicable in these cases:

$$C_x = \frac{25\times 10^9}{f^2 L_s}; \qquad L_x = \frac{25\times 10^9}{f^2 C_s}$$

where L is given in microhenries, C—in picofarads and f—in kilocycles. A considerable error of measurement is experienced in this method, owing to the influence of distributed capacitance C_L of the coil. This capacitance is usually of the order of some picofarads, but its exact value is unknown and the error, therefore, cannot be compensated for.

The measurement of capacitance can be carried out to a much higher degree of precision by the substitution method (Fig. 286*b*). Here, the generator is first adjusted to resonance with tuned circuit LC_x. After this, C_s is connected in place of C_x and the resonance is restored by varying the capacitance of C_s, the generator frequency being kept constant. In this method of measurement, there is no need to employ a standard inductance or to know the frequency value. Likewise, no error due to the influence of C_L is introduced in this case.

Fig. 285*c* shows a modification of the substitution method. At first, while C_x is cut out, the tuned circuit is adjusted to resonance, the condition of resonance corresponding to the case when the standard variable capacitance in the circuit has the value C_{s1}. After this, capacitor C_x is connected in parallel and resonance is restored by varying the value of standard capacitor to Cs_2. It is evident that $C_x = C_{s1} - C_{s2}$. If C_x is greater than the maximum value of C_s, capacitance C_x should be connected in series (Fig. 286*d*). In this case, C_{s2} is greater than C_{s1}, and the value of C_x is found from the following formula:

$$C_x = \frac{C_{s1}C_{s2}}{C_{s2} - C_{s1}} \, .$$

Fig. 286*e* shows a method of measurement of inductances L_x when the unknown inductances are very much smaller than the inductance of basic coil L. In this circuit, the resonance is first attained with the help of C_{s1}, while L_x is out of the circuit. Following this, L_x is connected in series with L and the resonance is restored by decreasing the capacitance to the value of C_{s2}. The value of L_x is then determined from the following formula:

$$L_x = \frac{25\times 10^9 (C_{s1} - C_{s2})}{f^2 C_{s1} C_{s2}}$$

where L_x is given in microhenries, the capacitances — in picofarads, and f — in kilocycles.

128. QUESTIONS AND PROBLEMS

1. Why are moving-iron instruments unsuitable for the measurement of high-frequency currents?
2. At what frequencies is it possible to use rectifier-type voltmeters?
3. How can a moving-coil meter be employed in the measurement of high-frequency currents?
4. Can a rectifier-type voltmeter be calibrated on direct current?
5. Why is it impossible to measure the anode voltage of a resistance-coupled amplifier valve by means of an ordinary voltmeter used for power-line measurements?
6. In what type of d.c. measurements, to be performed in electron-valve circuits, the 10,000 ohms-per-volt resistance of a voltmeter is not high enough?
7. What are the advantages and disadvantages of triode valve voltmeters?

8. The measurement range of a valve voltmeter may be extended by connecting a voltage divider, comprised of common ohmic resistors, at the input of the instrument. What is the shortcoming of such a method of range extension?

9. What types of instruments may be employed as high-frequency current-and-voltage indicators?

10. What precautions must be taken when measuring resistance values directly in the circuit of a radio receiver?

11. Describe the design of the simplest signal generator.

12. The sensitivity of an electron oscillograph is given as 3 mm/v. What does this mean?

13. What is the purpose served by the sweep-voltage generator in an oscillograph?

14. Draw the figure, displayed on the screen of an oscillograph, for the case when sinusoidal voltages on X and Y plates of the CRT have the following frequency ratio: $f_X : f_Y = 1:5$.

15. Why does an electron oscillograph need synchronisation?

16. What are the advantages and disadvantages of resonance wavemeters?

17. Describe the zero-beat frequency-measurement method.

18. A crystal-controlled heterodyne has a frequency of 2,000 kc. Can such a heterodyne be employed for the purpose of checking the calibration of a radio transmitter operating within the range of 800-1,600 kc?

19. Why is it that in an a.c. bridge the mere equality of resistance products of the opposite arms is insufficient to secure the state of balance?

20. Is it possible to measure various resistance in the circuit of a radio receiver directly by means of an a.c. bridge?

21. It is required to measure a capacitance of the order of 10 pf. What simplest method of measurement would you use in this case to obtain the highest accuracy?

CHAPTER XI

SEMICONDUCTOR DEVICES IN RADIO

129. GENERAL PROPERTIES OF SEMICONDUCTOR DIODES AND TRIODES

In recent years the art of semiconductor engineering has reached such a state of development that semiconductor devices are rapidly replacing the electron valve, previously considered the heart of modern radio. As a result, many types of radio equipment now employ semiconductors instead of valves.

The amplifying properties of semiconductor devices had for a long time remained unnoticed, although a typical semiconductor — the well-known crystal detector — has been employed in countless radio receivers from the very infancy of radio. These days the crystal detectors are more often called semiconductor rectifiers, or semiconductor diodes. Other types of semiconductors that have also been used in the past, and are still used at present, are the cuprous-oxide and selenium rectifiers.

As a matter of fact, in the earlier days all known types of semiconductors were considered only as rectifiers, although the cuprous-oxide and selenium varieties were not suitable for high-frequency rectifier applications, while the old-type crystal detectors did not provide sufficient stability. It was only after special types of semiconductors possessing amplifying properties were developed that the art of semiconductor engineering had really come to the fore.

Briefly referring to the history of semiconductors, it may be noted that there had been no well-founded semiconductor theory for a long time, although many cases of practical approach to the development of semiconductor circuits have been recorded. In this respect, the research carried out in 1922 by the Soviet engineer O.V. Losev is of great interest, because this work had shown that a semiconductor (a crystal) can be used as a generator of electromagnetic oscillations. The engineer had designed a special crystal receiver, the "crystaldyne", and the set had a much higher sensitivity than the usual (in those days) crystal receivers. The improved

sensitivity, in this case, was attributed to the generation of self-oscillations by the receiver. It is to be regretted that the discovered effect did not receive sufficient attention during the following years and the semiconductors continued to be regarded simply as rectifiers.

Only as late as 1948, first amplifying semiconductor devices that were put to practical use were developed in the United States of America. These devices are known now as crystal triodes or transistors. From that period on, the theory and the technology of the amplifying semiconductors, as well as of the non-amplifying semiconductors (crystal diodes), rapidly progressed. At present, the radio industry produces great quantities of semiconductor diodes and triodes of various types, the devices being successfully employed in different radio and electronic equipment.

In comparison with electron valves, semiconductor diodes and triodes are noted for the following advantages:

1. light weight and compactness;
2. no filament consumption;
3. longer service life (up to tens of thousands of hours);
4. greater mechanical strength (the devices are immune to shaking, blows and other types of mechanical overloads);
5. high efficiency of circuits using the devices, because the energy loss in semiconductors is very low;
6. very low voltages of power supplies feeding various semiconductor circuits (low-power amplifiers, oscillators, receivers, etc.).

Contrasted to the above-listed advantages, semiconductor devices have their own shortcomings, which are as follows:

1. considerable difference of parameters and characteristics of individual semiconductor devices of the same type and manufactured to the same specifications;
2. great dependence of the properties of the devices upon temperature, which does not allow us to operate these devices below —60° C or above +60° C (these limits can somewhat be extended in certain designs);
3. semiconductors age rather rapidly; i.e., the properties of the devices change with time, although the devices still remain serviceable;
4. higher level of internal noise than in an electron valve;
5. low frequency limit; as a rule, crystal triodes do not function on frequencies much higher than some dozens of megacycles;
6. the input resistance of crystal triodes is much lower than that of triode valves;
7. power limitations; semiconductor devices are manufactured in low-power ratings only.

(The last point refers only to semiconductor triodes. Semiconductor diodes, used for rectifier service in power supply instal-

lations, are designed for much higher ratings.—*Translator's note.*)

It should not be thought that the listed shortcomings of semiconductor devices are incurable. An extensive research is constantly conducted with the aim of improving the properties of such devices. Much has already been accomplished. Thus, new semiconductor materials are being developed. Superpower semiconductor rectifiers capable of carrying several thousand amperes of current are being designed.

Latest types of silicon triodes replacing germanium triodes will probably be capable of operation at temperatures as high as 250°C. The frequency limit is being conquered, too; special crystal triodes have been built to operate on frequencies as high as 1,000 mc. New crystal tetrodes are also designed for a considerably extended frequency range.

The limitation listed above pertain only to the present-day semiconductor devices, now widely used in radio and electronic equipment. The art of semiconductor engineering is very young and much can be expected from its further development.

130. ELECTRON-TYPE AND HOLE-TYPE CONDUCTIVITY IN SEMICONDUCTORS

Physical processes taking place in semiconductor diodes and triodes may be explained by the unique electrical conductivity properties of these devices. Extensive research conducted by Soviet and foreign physicists has revealed that there exist two basic types of semiconductors. Some semiconductors, such as oxides of aluminium, zinc or titanium, possess the so-called *electron-type conductivity*. This type of conductivity, possessed by metals under ordinary conditions, is also known as *n-type conductivity*. There is a great number of semifree electrons in semiconductors of the given type. The electrons are very loosely held by the atom-nuclei and perform a haphazard thermal motion between various atoms. But the atoms themselves are systematically distributed throughout the material and form the so-called crystal lattice. When a difference of potentials exists between the ends of a semiconductor, the electrons are forced to move in a definite direction, as shown in Fig. 287. Such unidirectional movement of electrons constitutes an electric current. Since the electrons are negatively-charged particles, moving as a stream from the negative terminal to the positive, the described type of conductivity may be called negative conductivity. The latter term is usually abbreviated as *n*-type conductivity, already mentioned above.

The other variety of semiconductors, such as protoxide of copper, selenium and many other substances, possess the so-called *hole-type conductivity*, otherwise known as *p-type conductivity* (*p*, in this

case, stands for "positive"). In these semiconductors the electric current is formed in a different way and can be considered as a shifting of positive charges. There are no semifree electrons in p-type semiconductors and, hence, electrons cannot move as they do in n-type semiconductors. Under the influence of the thermal effect, an atom of p-type semiconductor can lose one of the electrons most remote from the nucleus. Should this occur, the atom will be positively charged, the magnitude of such charge being numerically equal to the charge of the electron. It, however, would be wrong to assume that such an atom becomes an ion. There are, of course, certain media possessing ion-type conductivity. In such media, the movement of ions constitutes electric current (incidentally, the word "ion" means "traveller"). A p-type semiconductor, however, does not belong to this type of media and possesses a different system of shifting of electrical charges. The crystal lattice is sufficiently strong in a p-type semiconductor, i.e., in this case the atoms deprived of their electrons remain stationary.

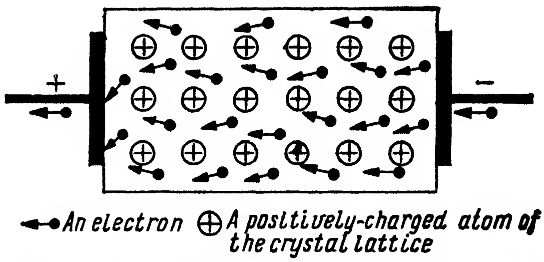


Fig. 287. The passage of electric current through a semiconductor possessing electron-type conductivity

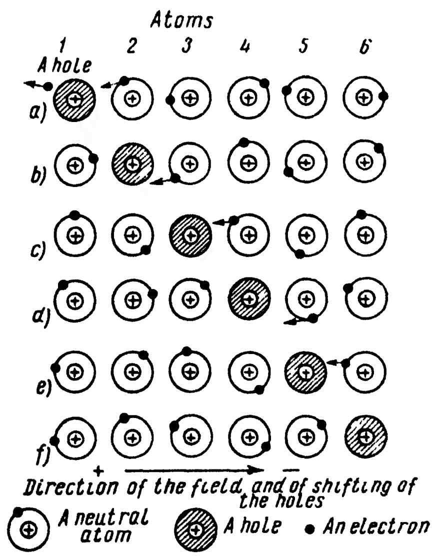


Fig. 288. The principle of hole-type conductivity

In a semiconductor, the absence of an electron from an atom makes the atom positively charged, the vacancy figuratively known as a "hole". This name indicates that the atom has been "robbed" of one of its electrons and a vacancy — the hole — is left in the atom for a free electron to "fall in". Theory and practice show that the holes behave as elementary positive charges. Hole-type conductivity indicates that not the electrons but the holes are shifted under the influence of potential difference applied to the semiconductor. This is equivalent to shifting of positive charges.

Fig. 288 illustrates the manner in which the holes move. Six atoms, lined up in a chain, are shown in this pictorial representation and are denoted by numbers 1, 2, 3, 4, 5 and 6. The chain is then regarded at different moments of time, these moments denoted as a, b, c, d, e and f. In order to understand the hole movement process, assume that at the initial moment "a" a hole has been formed in atom "1" because this atom was "robbed" of an electron (Fig. 288). Continuing a study of the same figure, we note the following.

The atom in which a hole has been thus created becomes positively charged and is capable of attracting an electron, "robbing" one of the neighbouring atoms of such an electron. If the semiconductor, in which the process is being studied, is subjected to the action of an electric field (i.e., if a potential difference is applied to the semiconductor), such electric field will have a tendency of moving the electrons in the positive direction — away from the negative terminal of the applied e.m.f. Because of this, the atom with the hole will in all probability "pull out" an electron from the nearest atom which is located closer to the negative terminal — from atom 2. This brings us to the next moment (moment "b") when atom 2 has just lost an electron to atom 1, thus "filling in" the hole in atom 1. The latter atom is no longer charged positively, because the necessary equilibrium has been restored between the positive and negative charges within the atom. However, atom 2, having lost an electron, becomes positively charged this time. This means that a hole has been formed in atom 2. The described process will now be repeated, but this time atom 2 will "steal" an electron from atom 3. As a result, a hole will be created in atom 3 (moment "c"), etc.

As the process continues, it becomes apparent that, after a certain lapse of time, the hole will be step-by-step transferred from atom 1 to atom 6. In other words, the positive electric charge, initially set up in atom 1, will be in the end conveyed to atom 6 (see moment "f" in Fig. 288).

As is evident from the above explanation, even in a semiconductor with hole-type conductivity it is actually a shifting of electrons that takes place. Such shifting is, however, much more limited than in a semiconductor with electron-type conductivity. In the hole-type semiconductor the electrons are simply transferred from one atom to a neighbouring one. What really characterises the hole-type semiconductor is that a movement of positive charges—of the holes—takes place in a direction opposite to the direction of electron movement.

Fig. 289 illustrates how electric current flows through a semiconductor with hole-type conductivity, the holes denoted by little circles and the electrons by points. When an electromotive force is applied to a semiconductor—if the semiconductor has the n-type conductivity (Fig. 287)—semifree electrons will flow through it,

as well as through the wires connecting it to the source of e.m.f. If, however, the semiconductor has a p-type conductivity (Fig. 289), the connecting wires will again carry a stream of electrons, but the current flowing through the semiconductor with have to be regarded as a flow of holes. The electrons enter the semiconductor from the negative terminal A and fill the holes reaching the same terminal, the "filling-in" process taking place right at the terminal. The process of uniting the electrons with the holes is known as *recombination*. The positive terminal B is reached by electrons from the neighbouring parts of the semiconductor, holes being formed in these parts, such holes moving from right to left.

Either n-type or p-type conductivity may be obtained in a given substance by adding different impurities to it. For instance, germanium, which is widely used in modern semiconductor devices, will possess n-type conductivity when such impurities as antimony or arsenic are added to it. This is attributed to the following. Each atom of antimony or arsenic interacts with the atoms of germanium and easily parts with one electron. As a result of such process, a great number of semifree electrons are formed. Impurities, whose atoms part with their electrons, are known as *donors*. If, however, germanium contains such impurities as indium or aluminium, the atoms of these substances will "steal" electrons from the atoms of the basic metal (i.e., the germanium), the process resulting in the formation of holes. All impurities accounting for the hole-type conductivity are called *acceptors*.

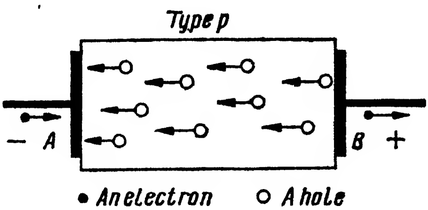


Fig. 289. The passage of electric current through a semiconductor possessing hole-type conductivity

It should be borne in mind that, in actuality, there exist no semiconductors with entirely electron-type or entirely hole-type conductivity. Every semiconductor usually possesses both types of conductivity; but one of the two types predominates. For instance, in a p-type semiconductor the main role is played by the hole-type conductivity, but in the same semiconductor the electron-type conductivity also helps the passing of electric current through the device.*

* The conductivity possessed by a pure semiconductor (i.e., a semiconductor free of impurities) is denoted by the letter i.

181. RECTIFICATION AT THE BOUNDARY OF TWO SEMICONDUCTORS

The boundary region between two semiconductors with opposite types of conductivity is known as *the p-n boundary*. This boundary possesses rectification properties. Fig. 290 illustrates a passage of current through the contact between two such semiconductors—for instance, between germanium p and germanium n.

Commencing our study of processes taking place at the p-n boundary, let us assume at first that the left-hand semiconductor possesses pure hole-type conductivity, while the right-hand one possesses pure electron-type conductivity. Also assume that a voltage from some external source is applied to the semiconductor in such a way, that the positive terminal is connected to semiconductor p, while the negative one is connected to semiconductor n (Fig. 290*a*). When the voltage polarity is as explained above, the electrons in semiconductor n will move towards boundary AB, the electron stream being met by an oppositely-directed stream of holes from semiconductor p. As the electrons and the holes meet, recombination takes place. Of course, only electrons move in the external wiring. The electrons flow in the direction of semiconductor n from the negative terminal of the voltage source. These electrons compensate for the electron loss occurring during the recombination process. The electrons leaving semiconductor p then flow in the direction of the positive pole of the voltage source, more and more new holes being formed in the given semiconductor. The described process goes on continuously and, therefore, electric current flows in the circuit. This current is known as the *forward current*. The p-n boundary offers but a very low resistance to this forward current. In other words, a considerable value of forward current can be obtained when a comparatively low external voltage is applied to the semiconductor.

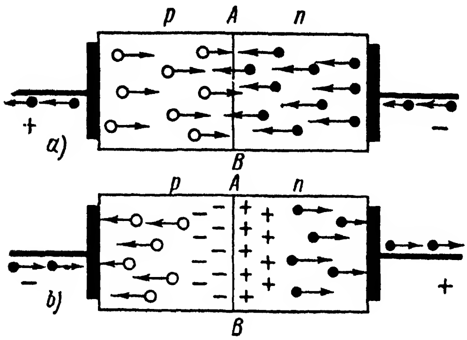


Fig. 290. The passage of electric current through two contacting semiconductors possessing different types of conductivity

An entirely different process will take place if the polarity of the voltage source is reversed (Fig. 290*b*). In this case, the electrons in semiconductor n will move in the direction of the positive terminal of the source, i.e., they move away from the p-n boundary. But meanwhile, a simultaneous movement of holes takes place in semiconductor p, the holes also moving away from the p-n boun-

dary. Upon reaching the left-hand electrode, the holes recombine with the electrons arriving from the wire connecting this electrode to the negative terminal of the voltage supply source.

Owing to the fact that a certain quantity of electrons have left semiconductor n, this semiconductor is charged positively, because positively-charged atoms remain in it. Apparently, these atoms are the atoms of that impurity (for instance, of arsenic) which, donating its electrons, had created the n-type conductivity in the germanium. Hence the positive charge of semiconductor n. While this is happening, semiconductor p will assume a negative charge, because the holes are leaving it (speaking strictly, electrons are entering this semiconductor to fill in the holes).

The described movement of the electrons and the holes in the opposite directions, when they are moving away from each other, can continue only for a short period of time. (This short-lived current resembles the current experienced during the charging of a capacitor.) Two opposite charges are created both sides of the pn boundary. As soon as the potential difference between these charges becomes equal to the e.m.f. E of the voltage source, the further movement of the charges will cease. The whole system will then resemble a charged capacitor. Under such conditions, the resistance of the p-n boundary is infinitely high and it is customary to say, in this case, that the so-called *barrier layer* has been formed at the boundary of two different semiconductors.

It should be noted that the above assumption of existence of the barrier layer of an infinitely high resistance at the p-n boundary does not actually hold true in reality. To be sure, the resistance will be very high, but still a certain amount of current will flow through it. The presence of current is attributed to the fact that, owing to the thermal processes, a comparatively small quantity of holes and semifree electrons is generated in every semiconductor. Because of this, small reverse current I_r always flows through the described circuit. The value of this current is usually much smaller than the value of forward current I_f. Thus, reverse resistance R_r is not infinitely high, although it is much higher than forward resistance R_f. At a certain, and quite low voltage value of the supply the reverse current reaches a steady value, after which this current no longer increases, even if the supply voltage is raised. In accordance with this, the steady value of the reverse current may be called the saturation value. The reason for the saturation is the limited quantity of carriers whose movement in a semiconductor creates the reverse current (a current of electrons in the p region, and a current of holes in the n region). If the ambient temperature, at which the semiconductor operates, is increased, the quantity of secondary carriers also increases. This causes an increase of the reverse current and, hence, a decrease of the reverse resistance of the semiconductor.

Owing to the presence of reverse voltage, the *pn* boundary to a certain extent resembles a leaky capacitor. When such a capacitor is being charged, at first, during a short period of time, a considerable charging current will flow, following which only a small leakage current will be passing through the circuit. There is, however, a difference between the leakage-current phenomenon and the phenomenon of the reverse current in a semiconductor. The leakage current is proportional to the voltage applied to the capacitor, while the reverse current passing through the *p-n* boundary depends but slightly upon the e.m.f. across the semiconductor. It should be noted that the *p-n* boundary offers some capacitance also to the forward voltage, but in this case the capacitance is shunted by the low resistance of the boundary.

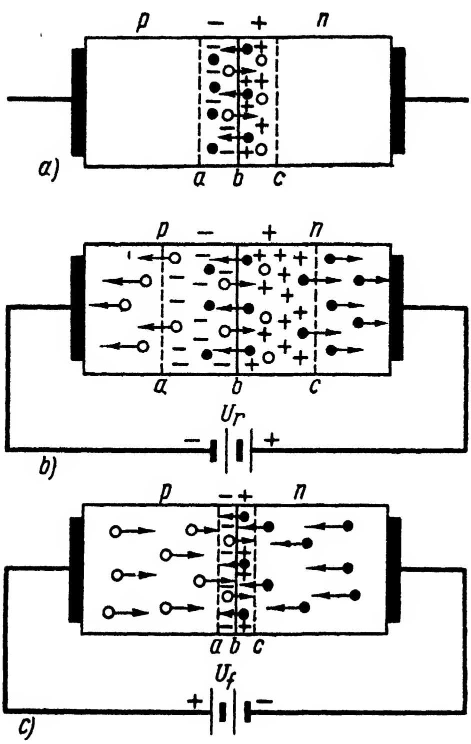


Fig. 291. The formation of barrier layer *(ac)* between two contacting semiconductors *(a)*. The broadening of the barrier layer under the influence of the reverse voltage *(b)*. The narrowing of the barrier layer under the influence of the forward voltage *(c)*

Thus, a contact between two semiconductors possessing opposite types of conductivity results in different values of resistance offered to the forward and reverse currents. Obviously, this effect can be utilised for the purposes of rectification.

A more thorough investigation of the processes taking place at the *p-n* boundary reveals that a barrier layer is formed at the contact surface of different semiconductors even when no external voltage is applied. To show that such is the case, let us examine a contact between *p*-germanium and *n*-germanium (Fig. 291*a*). There are many semifree electrons in *n*-germanium (these electrons are referred to as basic carriers), and few holes (secondary carriers). On the other hand, *p*-germanium possesses many holes and few semifree electrons. As a result of the haphazard thermal movement of the carriers, a diffusion takes place and the carriers are transferred from one semiconductor to the other. Electrons diffuse from *n*-germanium to *p*-

germanium, while the holes diffuse in the opposite direction. Opposite tied charges are created both sides of the boundary, a positive charge — in n-germanium, a negative charge — in p-germanium. The so-called contact potential difference is set up between these charges and the resulting electric field will oppose any further diffusion of the carriers. Still, even in the stabilised condition, the boundary will be passed by certain (although small) quantities of electrons and holes in both directions. This is attributed to the haphazard thermal movement of the carriers, some of which always have a sufficient energy to overcome the contact potential difference. Thus, the state of p-n boundary, shown in Fig. 291*a*, is a state of dynamic equilibrium.

Boundary layers *ab* and *bc* possess a decreased quantity of basic carriers and their resistance is higher than the resistance of the remaining part of the semiconductor. The combination of these layers, i.e., the whole of the *ac* sector, is the barrier layer referred to above. If now an external voltage source is switched on and its positive terminal is connected to n-germanium, while its negative terminal is connected to p-germanium (Fig. 291*b*), the external field created by the supply at the *pn* boundary will be added to the internal field set up by the contact potential difference. The resultant field will be intensified and a greater obstruction to the transformer of the carriers across the boundary will be experienced. Besides this, due to the influence of the external potential difference, the basic carriers in both semiconductors will be moving away from the boundary. As a result, the barrier layer will become thicker and its resistance will increase.

However, if the external voltage source is made to send current in the forward direction (Fig. 291*c*), the field set up by this voltage will be opposed to the internal field. The resultant field at the p-n boundary will then be weakened and the passage of the basic carriers across the boundary will be facilitated. These carriers, filling in the barrier layer, will decrease the thickness of the layer. The resistance of the layer will sharply drop and, at a certain value of the applied voltage, will disappear altogether.

132. SEMICONDUCTOR DIODES

The one-way conductivity of a semiconductor diode is clearly evidenced by its *volt-ampere characteristic*, which is given by a curve representing the relation between current and voltage. An example of such characteristic is shown in Fig. 292. The characteristic pertains to a low-power semiconductor diode and shows that a considerable forward current (several dozen milliamperes) is obtained even when the voltage applied in the forward direction is as low as a few fractions of a volt. It follows from this, that the forward resis-

tance in a semiconductor diode is very low — a few dozen ohms. In higher-power semiconductor diodes the forward current can be as high as hundreds of milliamperes and even higher when a similar low voltage is applied to the diode. In such diodes, the boundary resistance R_{np} is accordingly decreased to a few ohms and even lower.

That part of the characteristic which represents the reverse current (usually much lower than the forward current) is generally plotted to a different scale than the forward characteristic (Fig. 292). As may be seen, even when the voltage applied in the opposite direction is equal to several dozen volts, the reverse current does not exceed a fraction of one milliampere. This signifies that the reverse resistance, in the given case, is expressed in several dozen kilohms. The value of such resistance R_r ranges from a few kilohms to hundreds of kilohms in various types of semiconductor diodes.

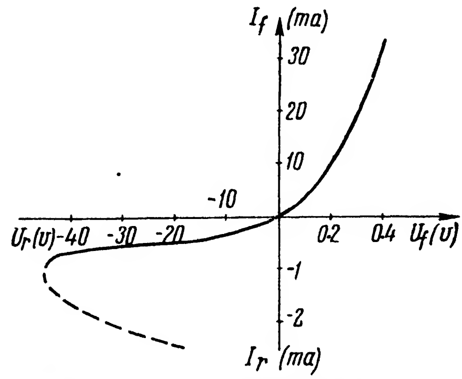


Fig. 292. The volt-ampere characteristic of a semiconductor diode

Of course, a voltage applied in the reverse direction should not be made too high. If this voltage exceeds the normal U_r rating of a given diode, the barrier layer of the diode will be punctured. Should this occur, the resistance of the barrier layer will sharply decrease, the reverse current will rise to a high value, and voltage will suddenly drop. The dotted part of the characteristic given in Fig. 292 shows how the characteristic will behave when a puncture of the barrier layer occurs.

In some cases semiconductor diodes are characterised by the value of rectification factor k_r. This is the ratio of the forward current to the reverse current, or the ratio of the reverse resistance to the forward resistance, for identical values of the forward and reverse voltages. The value of k_r is found from the following formula:

$$k_r = \frac{I_f}{I_r} = \frac{R_r}{R_f}\,.$$

It should be noted, however, that owing to non-linearity of the volt-ampere characteristic, the value of k_r is very unsteady. Because of this, modern design practice does not use the rectification factor, preference being given to denoting the values of I_f and I_r for definite forward and inverse voltages. An alternative way is to give the characteristic of the diode. This gives a more complete idea of its rectification properties.

Cuprous-oxide and selenium rectifiers, discussed under Chapter V, are widely used as rectifiers in power supplies of different types of radio equipment. The operation of these semiconductors is also based on the principle of barrier-layer formation at the p-n boundary. The copper protoxide, used in the cuprous rectifiers, covers a copper electrode. This protoxide has a hole-type conductivity. However, during the manufacturing process, a layer is formed between the protoxide and the copper. This layer differs in its chemical composition from the protoxide and possesses electron-type conductivity. The rectification takes place at the boundary of this layer and the protoxide (Fig. 293).

In selenium rectifiers, the selenium, placed on an aluminium or steel base, has a hole-type conductivity. The fusible metal, covering the selenium, contains cadmium. Combining chemically with the selenium, the cadmium forms a semiconducting layer possessing an n-type conductivity. The boundary between this layer and the selenium is the p-n boundary (Fig. 294a). For operation at higher voltages (30 volts and higher), special selenium diodes (type T) are manufactured, in which the layer of electron-type semiconductor material is located between the selenium and the aluminium base. In these diodes, not the cadmium-containing plate but aluminium foil is used as the second electrode. In this case, the basic aluminium plate serves as a cathode and not as anode, differing in this respect from the usual diodes (Fig. 294b).

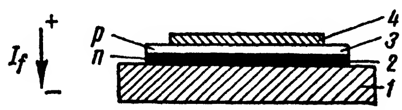


Fig. 293. The design principle of a cuprous-oxide diode: 1 — copper; 2 — barrier layer; 3 — copper protoxide; 4 — upper electrode

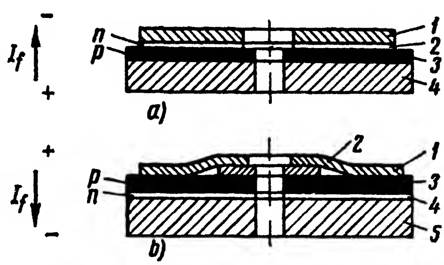


Fig. 294. The design principles of selenium diodes: a) the usual type of diode: 1 — the cathode alloy; 2 — barrier layer; 3 — selenium; 4 — aluminium or steel (anode); b) type T diode: 1 — aluminium foil; 2 — mica washer; 3 — selenium; 4 — barrier layer; 5 — aluminium (cathode)

A prominent role is played in the modern field of radio and electronics by germanium and silicon diodes, whose applications are very numerous and still increasing (although other types of semiconductor diodes also exist). There are two general types of germanium and silicon diodes, namely — *junction diodes and point-contact diodes*. The junction diodes are also called *power diodes* and are used in power rectifier circuits. The p-n boundary has quite a large surface in these diodes, which accounts for a considerable

power-handling capability of these rectifiers. However, junction diodes possess high internal capacitance (20 pf and greater) and their use is, therefore, confined to circuits operating on frequencies below 50 kc.

Fig. 295 shows the design and external view of a junction diode of Soviet manufacture. This diode, known as type ДГ-Ц, represents a little cylindrical metal container enclosing a germanium plate (crystal) which possesses an electron-type conductivity. The plate is soldered to a metal base, the latter contacting with the body of the container. The indium impurity is introduced into the other side of the germanium, forming a sphere with hole-type conductivity in the basic metal. A terminal soldered to this sphere passes through a glass insulator embedded into the body of the diode. The external diode terminals are ribbon-shaped. The dimensions of the diode are given in Fig. 295.

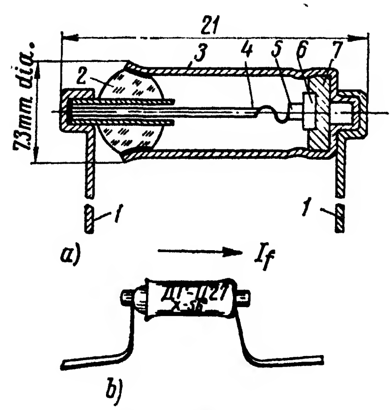


Fig. 295. The design principle and external view of junction diodes type ДГ-Ц: *1* — terminals; *2* — glass; *3* — body; *4* — pick-up element; *5* — indium; *6* — germanium; *7* — crystal holder

Below are listed the basic parameters of rectifying semiconductor diodes, as well as the designation of such parameters.

1. Maximum permissible rectified current I_{-max}. (This current represents the average value or the d.c. component of the rectified pulsating current.)

2. Forward voltage drop U_f across the diode when it is passing current I_{-max}.

3. Maximum permissible reverse voltage $U_{r\ max}$.

4. Maximum reverse current $I_{r\ max}$, flowing when $U_{r\ max}$ is present.

These parameters, as well as other parameters of a crystal diode, pertain to the operation of the device within a temperature range extending from 15° to 25° C.

Up to the present, the electronic industry has been producing seven types of junction germanium diodes, the types ranging from ДГ-Ц21 to ДГ-Ц27. The first four types are characterised as follows:

$I_{-max} = 300$ ma;
U_f—not over 0.5 v;
$I_{r\ max}$—not over 0.5.ma;
$U_{r\ max}$—50, 100, 150 and 200 volts, respectively.

The last three types of diodes in the above series, namely diodes ДГ-Ц25, ДГ-Ц26 and ДГ-Ц27, have the following characteristics:

$I_{= max} = 100$ ma;
U_f—not over 0.3.v;
$I_{r max}$—not over 0.3 ma;
$U_{r max}$—300, 350 and 400 volts, respectively.

The reverse breakdown voltage of all these diodes is at least 1.5 $U_{r max}$. The diodes will withstand the passage of 25-ampere current through them, if the current flow does not last longer than 0.1 sec. All these types of crystal diodes may be operated at ambient temperatures within the range extending from −60° C to +70° C. It should be noted, however, that the rectifying properties of the diodes are impaired when the temperature is considerably higher or lower than the normal temperature of +20° C, at which the diodes give the best performance. If a crystal diode of this type has to be operated at temperatures higher than +20° C, the value of the a.c. voltage being rectified or the value of the rectified current should be reduced in order to prevent overloading of the diode. It is not advisable to operate these diodes when the ambient temperature exceeds +70°C as the diode is likely to be damaged under such working conditions.

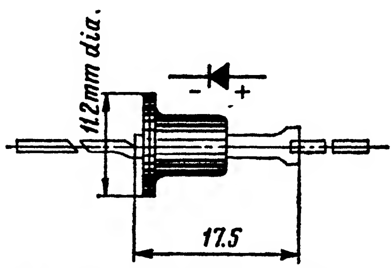


Fig. 296. Junction germanium diodes type Д7

The latest crystal diodes produced by the Soviet electronic industry are represented by a group consisting of seven types ranging from Д7А to Д7Ж. The parameters of these diodes are similar to those of the ДГ-Ц series, but the new diodes are hermetic and are of smaller dimensions (Fig. 296). The reverse current in the first four of the new diodes is also smaller and does not exceed 0.25 ma.

The industry is also producing silicon junction diodes in the type range from Д201А to Д201Ж. These diodes are made of *n*-type silicon, aluminium being fused into the basic metal to form a *p*-region. Various types of such silicon diodes are characterised as follows:

$I_{= max}$—from 200 to 400 ma;
U_f —from 1.5 to 2 v;
$U_{r max}$—from 25 to 250 v;
$I_{r max}$—not higher than 0.5 ma.

Junction diodes of both the germanium and silicon varieties may be used in any type of rectifier circuits discussed under Chapter V.

Point-contact diodes are designed to handle smaller power than junction diodes. However, the point-contact diodes possess a much

lower capacitance (not over 1 pf) and can, therefore, be employed on high frequencies up to 150 mc.

Fig. 297 shows the design principle of a point-contact diode. Here, a thin tungsten wire is brought into contact with a germanium plate. The germanium used has an n-type conductivity, but during the so-called forming process a small sector with p-type conductivity is developed in the basic metal near the point where it makes the contact with the wire. Thus, in a point-contact diode rectification takes place at the n-p boundary between two different semiconductors, just as it does in a junction diode. Hence, there is no radical difference between the point-contact diodes and junction diodes. But, because the area of the n-p boundary is much smaller in a point-contact diode, this diode can handle only smaller currents than a junction diode, although it possesses the advantage of lower capacitance over the junction device.

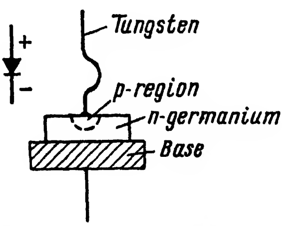


Fig. 297. The design principle of a point-contact diode

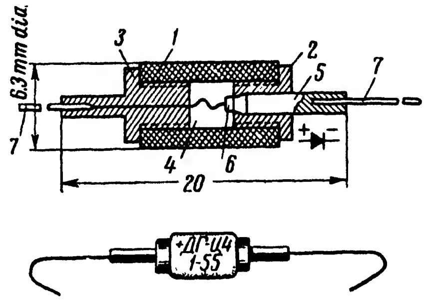


Fig. 298. The design principle and external view of point-contact diodes type ДГ-Ц: *1* — ceramic cartridge; *2* and *3* — metal flanges; *4* — contact spring; *5* — crystal holder; *6* — germanium; *7* — terminals

Point-contact diodes produced by Soviet industry carry similar marking as the junction diodes already discussed above. Thus, there is a ДГ-Ц series of junction diodes and a ДГ-Ц series of point-contact diodes. Fig. 298 shows the design and external view of a point-contact diode of the ДГ-Ц series. The germanium crystal has a polished surface with an area of about 1 sq mm. The tungsten wire, approximately 0.1 mm thick, has a sharpened end which bears upon the germanium. The area of contact between the two metals comprises only a few square microns. The diode is enclosed in a ceramic cartridge provided with metal flanges. Thin flexible wires ("pigtails") are used for the terminals. The parameters of point-contact diodes are designated in the same way as the parameters of junction diodes (see above). There is also an additional designation in a point-contact diode, namely — forward current I_f that the device will pass when a voltage of 1 volt is applied to it.

These point-contact diodes are designed for values of $I_{-\,max}$ up to 16 ma (in case of ДГ-Ц8 up to 24 ma). $U_{r\,max}$ and $I_{r\,max}$ values

range from 30 to 200 v and from 0.06 to 1 ma for various diodes of this series. The forward current of these point-contact diodes may be from 1 to 10 ma. The 1-sec current surge may be as high as 300 ma (in case of ДГ-Ц8—up to 500 ma). The normal ambient temperature for these diodes ranges from +15° C to +25° C, although they can safely operate within the limits extending from —50° up to +70° C. All ДГ-Ц diodes may be used as rectifiers in measuring instruments in the role of detectors, limiters, etc.

The latest types of point-contact germanium diodes are enclosed in metal-glass containers (Fig. 299). This new diode series, known as

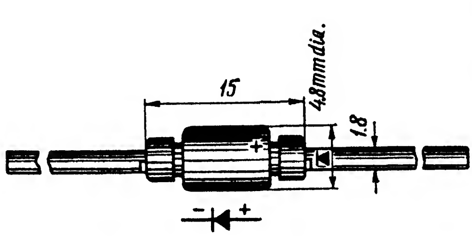


Fig. 299. Point-contact germanium diodes type Д2

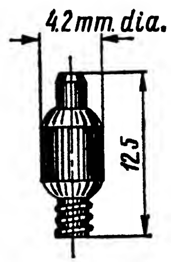


Fig. 300. External view of point-contact mixer diodes type ДГ-С

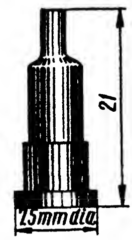


Fig. 301. External view of point-contact diodes ДК-И and ДК-С

the Д2 series, ranges from Д2А to Д2Ж. Diodes beloging to this series have similar parameters as diodes of the ДГ-Ц series.

Besides the diodes intended only for rectifier duty, special germanium diodes, intended for the mixer service, are also manufactured and range from type ДГ-С1 to type ДГ-С4. These diodes are used for frequency-conversion purposes in superheterodyne radio receivers designed for centimetre waves. The diodes are of the point-contact version. Constructionally, they represent ceramic cartridges provided with stub terminals for connection into the circuit (Fig. 300).

Silicon diodes are also being produced for operation on centimetre waves. These also are point-contact versions, and are successfully employed in receiving and measuring equipment designed to operate on extremely high frequencies. In comparison with the germaninm diode, the silicon diode has a higher sensitivity, i.e., its volt-ampere characteristic is steeper on forward current. Hence, the silicon diode will pass larger currents at lower voltages. In other words, the silicon diode possesses lower resistance in the forward direction of current flow. Besides this, the silicon diode — again in comparison with the germanium diode — has a smaller capacitance, a lesser dependence of operation upon the ambient temperature, and a lower level of internal noise. The shortcoming of the silicon diode is its low maximum permissible reverse voltage, which does not exceed a few dozen volts.

Silicon diodes are produced in the following three series:

1. ДК-В series, extending from ДК-В1 to ДК-В7, designed for detection work in receivers;

2. ДК-И series, extending from ДК-И1 to ДК-И2, designed for use in measuring equipment;

3. ДК-С series, ranging from ДК-С1 to ДК-С5, designed for frequency conversion duty.

Constructionally, diodes of the ДК-В series, ranging from ДК-В1 to ДК-В4, are similar to the diode shown in Fig. 300. The remaining diodes of this group are enclosed in ceramic cartridges and are designed as shown in Fig. 301. All point-type germanium and silicon diodes intended for operation on centimetre waves will also give a good account of themselves on lower frequencies.

133. SEMICONDUCTOR TRIODES (TRANSISTORS)

Semiconductor triodes, which shall be referred to in the further discussion as transistors (their accepted name in various countries), are at present widely used in various amplifiers, receivers, oscillators and countless electronic devices. Like the semiconductor diodes, the transistors, too, come in two types—junction type and point-contact type. The principle of operation of transistors can be most conveniently explained referring to the junction transistor.

Fig. 302 gives a block diagram of the junction transistor. The device is nothing but a germanium plate (although other semiconductors may be used in place of the germanium) in which three zones of different conductivity have been created. As an example, we take a transistor in which the middle zone has a hole-type conductivity, while the two end zones possess conductivity of the electron type. A transistor like this is called a *npn transistor*. Transistors of the *pnp type* are also used in radio engineering. Physical processes taking place in both types of transistors are analogous.

The middle zone of the transistor is called *the base* (although it is sometimes referred to as *the control electrode)*. One of the end zones is called *the emitter*, the other one—*the collector*. As may be seen from the drawing, there are two *np* boundaries in a transistor, one of which is known as *the emitter junction*, the other one—*the collector junction*. The distance between these two junctions is usually very small—from 10 to 20 microns. Thus, the middle zone, in actuality, is a very thin layer. This is the most important factor in the function of the transistor.

In transistor circuit diagrams, letters *b*, *e* and *c* are used to designate, respectively, the base, the emitter and the collector of a transistor. Constructionally, each transistor is provided with separate terminals connecting the three electrodes of the device to external wiring. Currents flowing through the three terminals are designated, res-

pectively, as currents I_b, I_e and I_c. Voltages appearing between the emitter and the base and between the collector and the base are designated as U_e and U_c. Standard schematic representations of the *pnp* and *npn* transistors are shown in Fig. 303.

The transistor is connected into its associated circuit in such a way, that two circuits are formed. *The input* or the *controlling circuit* is analogous to the grid circuit of a triode valve. The *output* or the *controlled* circuit resembles the anode circuit of an electron valve. The oscillations to be amplified are fed to the input circuit of the transistor, its output circuit terminating in a load resistance

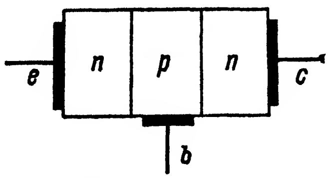


Fig. 302. The design principle of a junction transistor

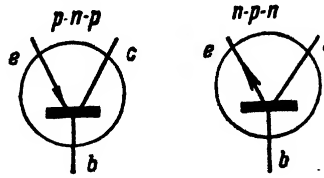


Fig. 303. Schematic representation of transistors in circuit diagrams

across which the amplified voltage appears. Index 1 is attached to the input circuit designations and index 2 to the designations of the output circuit. Thus, currents and voltage in these circuits are respectively designated as I_1, U_1, I_2 and U_2.

Although the transistors, just as the electron valve, are designed to amplify electrical oscillations, the physical processes taking place in the two devices are quite different. Therefore, when speaking of similarity between valves and transistors, it should be understood that they are only analogous, and not identical in their functions. Keeping this in mind, the emitter may be compared to the cathode of a valve, the collector—to the anode, and the base—to the grid.

There may be three general versions of electron-valve amplifier stages. As we know, it is a general practice to employ *earthed-cathode* stages, which may be regarded as stages with a *common cathode.* Various amplifier stages that we have already studied employ just such a type of circuit arrangement. The cathode of the valve in a stage of this kind serves as the branching-off point of the grid and anode circuits. The cathode, as a rule, is earthed. There are, however, certain alternatives. Thus, in Sec 79 of Chapter VII, dealing with the application of negative feedback in low-frequency amplifiers, there is described a circuit of a cathode follower operating with an earthed anode (earthed, as far as a.c. voltage is concerned), the load being connected into the cathode wire. In such a circuit, not the cathode but the anode is the branching-off point of alternating currents of the grid and anode. Thus, the cathode follower represents an amplifier stage with a *common anode* (or *earthed anode).*

Another alternative is the ultra-high-frequency circuit proposed by the Soviet scientist M.A. Bonch-Bruyevich. This is an amplifier circuit in which the grid is earthed. Such *earthed-grid* circuit may be called a *common-grid* circuit.

Just as in the above-described cases, amplifier stages employing transistors are also classified as *common-emitter*, *common-collector* and *common-base* stages. The alternative terminology is: *earthed-emitter*, *earthed-collector* and *earthed-base* circuits. All these circuits and the analogous valve circuits are shown in Fig. 304 for facility of comparison.

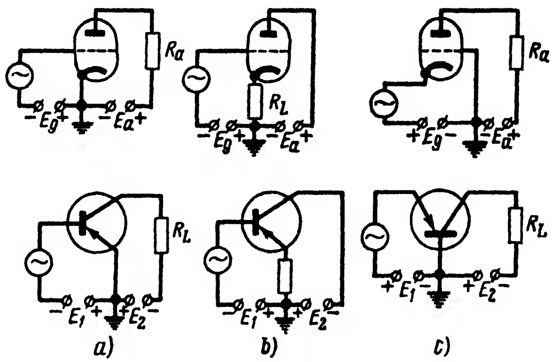


Fig. 304. Comparison of amplifier stages employing electron valves and type *p-n-p* transistors: *a)* circuits with earthed cathode and earthed emitter; *b)* circuits with earthed anode and earthed collector; *c)* circuits with earthed grid and earthed base

Various types of amplifier stages possess different properties, but the general principle of amplification is, of course, the same. Therefore, we shall analyse in detail the operation of an earthed-base transistor. As a matter of fact, this circuit is widely used and is encountered in practice much more often than its analogue—the earthed-grid electron valve amplifier.

In the earthed-base transistor amplifier, the input is the emitter circuit and the output—the collector circuit. Therefore, in such a circuit $I_1 = I_e$; $U_1 = U_e$; $I_2 = I_c$ and $U_2 = U_c$. In our further discussion, we shall use indexes 1 and 2 without any explanation, whenever the values carrying such indexes will pertain to earthed-base stages, denoting emitter and collector circuits, respectively.

Let us first analyse the operation of a transistor (for example of the *n-p-n* type) under the condition of static operation, when only d.c. supplies, generating voltages E_1 and E_2, are connected in the circuits of emitter and collector. The common-base circuit for this case is shown in Fig. 305. The polarity of supply voltages must be such that the emitter boundary works in the forward direction,

while the collector boundary works in the reverse direction. The resistance of the emitter boundary is small and, therefore, the value of voltage E_1 need only be a fraction of one volt in order to secure a normal value of current in the emitter. On the other hand, the resistance of the collector boundary is high and, hence, voltage E_2 is always considerably higher than voltage E_1 and usually ranges from some volts to several dozens of volts.

If the collector circuit is open, the volt-ampere characteristic of the emitter boundary will be represented by the right-hand part of the characteristic pertaining to a diode and shown in Fig. 292. And if the emitter circuit is open, the volt-ampere characteristic of the collector boundary will not differ from the left-hand part of the same diode characteristic, corresponding to the reverse voltage.

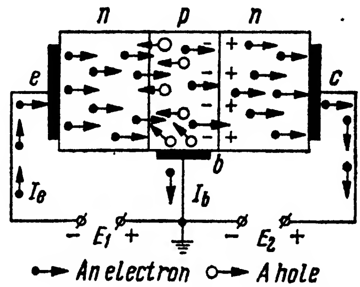


Fig. 305. Paths travelled by electrons and holes in a n-p-n transistor

Now assume that both circuits are closed. This will immediately give rise to a new effect; the emitter current will exert a strong influence upon the current of the collector, and *the greater the emitter current, the greater will be the collector current*. Thus, the emitter current controls the collector current (compare this to a valve stage, where the grid voltage controls the anode current). The amplification of electrical oscillations in a transistor is obtained on the basis of the effect just described, which calls for the following explanation.

Emitter current electrons, having passed through the emitter boundary, continue to move through the base and penetrate into the region of the collector boundary, which increases the collector current. The collector boundary works with the reverse voltage. Considerable charges, denoted by "+" and "—" in Fig. 305, are developed in the region of collector boundary. A positive charge is set up to the right of the boundary, in the n section, and a negative charge—to the left of the boundary, in the p section. An electric field is set up between these charges. This field helps the electrons (which have reached this point in the base after passing through the emitter boundary) to pass through the collector boundary. If the thickness of the base is sufficiently small, most of the electrons, in passing through the thin layer of the base, do not find enough time to recombine with the holes and successfully reach the collector boundary. Only a small quantity of electrons which have passed through the emitter boundary recombine with the holes, and give rise to the current I_b flowing through the base wire. However, if the thickness of the base is considerable, most electrons of the emitter current are caught by the holes in the base and will not reach the collector

boundary. In this case, the collector current will be unaffected by the emitter current electrons, and the latter will contribute only to an increase of the base current.

The increase of collector current on account of the emitter current is actually explained by the fact that emitter current electrons reach the collector boundary and lower its resistance. If $I_e = 0$, the collector boundary region has high resistance, because the basic carriers of the electric charges are moving away from this boundary in different directions. In the n region the electrons move to the positive electrode, while in the middle p region the holes move to the base electrode. A region deprived of basic carriers is created both sides of the p-n boundary. As a result of this, the collector boundary possesses high resistance and passes only a small reverse current. This current is formed due to shifting of secondary carriers towards each other, these carriers being the electrons from the p region and holes from the n region.

When the emitter current flows, the p region is reached from the direction of the emitter boundary by those electrons which are extra, non-essential, carriers for that region. Failing to recombine with the holes in the p region, these electrons reach the region of the collector boundary and decrease its resistance. The greater the emitter current, the greater will be the quantity of such extra electrons reaching the collector boundary and the lower will become the resistance of the boundary. When this happens, the collector current will correspondingly increase.

The name "emitter" signifies that electrons are emitted from the left-hand n zone to the base. This effect is often referred to as "*injection*". When speaking of transistors, it is advisable to use the term "injection" in order to distinguish between the effect of emission taking place in the electron valve and the effect taking place in the transistor. Evidently, injection in a transistor is quite different from emission of electrons by a heated cathode in a vacuum or in a rarefied gas.

The above examination of the processes taking place in a transistor has revealed a similarity between transistors and valve triodes, but it has also revealed that there are certain points of difference between the two types of devices. In a valve, the grid controls the anode current by means of an electric field. Grid voltage is the controlling factor in this case. As the grid voltage changes, it varies the intensity of the field existing between the grid and the cathode. In accordance with this, a greater or lesser amount of electrons emitted by the cathode (and located in the space charge around it) move to the anode through the grid. The value of anode current is determined by the grid voltage, but the grid current may be totally absent. As we know from our studies of electron valve circuits, in most cases the grid current is quite useless and even undesirable. Because of this, in most cases valve amplifiers operate without

grid currents, the latter being prevented by applying a certain amount of bias voltage to the grid. Now, referring to transistors, we see the following. When voltage varies in the emitter circuit, the emitter current is changed. This varies the amount of electrons penetrating from the emitter into the region of collector boundary, and, in accordance with this, the collector current varies. However, a part of the emitter current is always branched off into the base wire. Hence, a transistor cannot operate without the base current. Because of this, the operation of transistors resembles the operation of positively-biased electron valves.

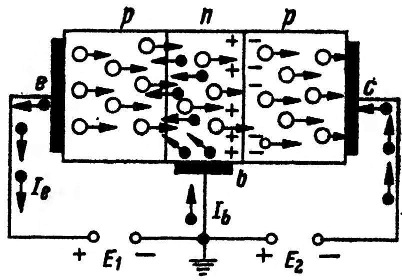


Fig. 306. Paths travelled by electrons and holes in a *p-n-p* transistor

Above we have studied physical processes taking place in *n-p-n* transistors. Similar processes take place in *p-n-p* transistors, but in this case the functions of electrons and holes are interchanged. The polarity of voltages and direction of currents are reversed. Fig. 306 shows how the *p-n-p* transistor is connected. In contrast to the *n-p-n* transistor, in which the electrons penetrate to the region of the collector boundary across the base, in the *p-n-p* transistors, holes, and not electrons, travel the same path. The holes are the secondary carriers with respect to the base, and they lower the resistance of the collector boundary. As the emitter current increases, a greater number of holes get through to the collector boundary. This causes a decrease of the resistance of the given boundary and increases the collector current. Junction transistors of the Soviet make belong to *p-n-p* type.

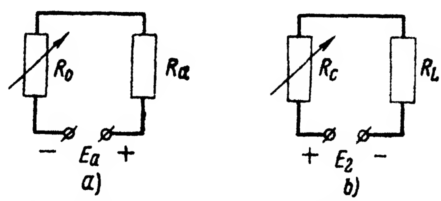


Fig. 307. Equivalent circuits of two amplifier stages, one of which employs an electron valve *(a)* and the other —a transistor *(b)*

The operation of an amplifier stage employing a transistor is similar to the amplification of oscillations by means of an electron valve. In the amplifier stage using a triode valve, voltage developed by the anode supply is divided between anode resistance R_0 of the valve and resistance R_a of the load. The equivalent of the anode circuit of such a stage is given in Fig. 307*a*.

If a.c. voltage is fed to the grid of the valve, resistance R_0 will change. The higher the negative voltage of the grid, the greater will be the value of R_0. When the grid voltage is sufficiently high to cut off the anode current, R_0 becomes infinity. Conversely, when grid

voltage becomes more positive in relation to the cathode, the value of R_0 decreases. Since the value of R_a remains constant as R_0 changes, voltage E_a of the supply source is constantly redistributed between R_0 and R_a while alternating voltage is applied to the grid of the valve. Considerable voltage changes are produced across R_a, although R_0 changes are caused by comparatively small grid-voltage changes.

The above function of the valve circuit provides, as we may recall, the amplification of alternating voltage. Besides, a valve stage also gives current and power amplification. Even when we take a case in which the alternating voltage across R_a is equal to the alternating voltage acting in the grid circuit, power amplification would still take place because the alternating current is greater in the anode circuit than in the grid circuit. But in most cases the grid current is so small that it may be considered zero. Consequently, the power in the grid circuit is very much lower than in the anode circuit, i.e., we have a case of tremendous amplification of power. The amplified power is obtained by virtue of conversion of anode-supply d.c. power into a.c. power, such conversion taking place in the anode circuit of the stage.

Let us now take an amplifier stage employing a transistor. In this stage, voltage E_2 obtained from the collector power supply is divided between resistance R_L of the load and the internal resistance offered by the transistor to the direct current of the collector. This resistance can be considered to be wholly consisting of the resistance offered by the collector boundary to direct current. Actually, the small resistance, R_b, of the base ought to be added to this resistance. However, R_b is so low that it may be disregarded. Thus, the equivalent of the collector circuit looks as shown in Fig. 307*b*. This equivalent circuit is quite analogous to the circuit shown in Fig. 307*a*.

When a source of oscillatory voltage is connected into the emitter circuit, the emitter current will vary in accordance with voltage changes. This, in its turn, results in corresponding resistance changes in the collector boundary. Then, voltage E_2 of the source will be repeatedly redistributed between R_c and R_L. The alternating voltage across the load resistor is dozens or hundreds of times as large as the alternating voltage in the emitter circuit. In a junction transistor, collector current changes are approximately equal to the emitter current changes. Hence, no current amplification takes place in the earthed-base circuit; but the power is amplified by the same number of times as the voltage. The amplified power is a part of d.c. power delivered by supply E_2. As will be shown later, earthed-emitter and earthed-collector circuits provide considerable amplification of current also.

The so-called *static coefficient of current amplification* α is the most important parameter of a transistor. This coefficient is the ratio of a collector current change ΔI_c to the emitter current

change ΔI_e, collector voltage U_c being maintained at a constant value (it should be kept in mind that collector current changes are caused by emitter current changes). This relation may be expressed mathematically as follows:

$$\alpha = \frac{\Delta I_c}{\Delta I_e}; \qquad U_c = \text{const.}$$

As far as junction transistors are concerned, α is always less than 1. When the transistor is operated on low frequencies, the value of α varies usually between 0.8 and 0.97, but in some cases may be slightly higher, although always less than unity. As the frequency increases, α becomes smaller. This is explained by a certain amount of inertia possessed by the carriers during their travel through the base, and is also explained by the undesirable shunting effect of the internal capacitance of the collector boundary. This capacitance is comparatively high in junction transistors and reaches several dozen picofarads. The given capacitance increases with an increase of the boundary area, and, therefore, is most significant in high-power transistors. Modern junction transistors are designed to operate on frequencies not exceeding a few megacycles. There is no doubt that further research in the field of transistors will lead to the development of junction transistors possessing higher frequency limits. The work performed at present has already given very promising results in this connection.

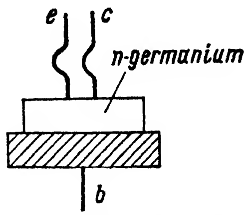


Fig. 308. The design principle of a point-contact transistor

The construction principle of point-contact transistors is schematically shown in Fig. 308. Here, a germanium plate, usually possessing n-type conductivity, is secured to a metal base provided with a terminal. Two thin sharpened wires, made of tungsten or some other metal, are made to bear upon the germanium plate. One of these wires is the emitter terminal, while the other is the terminal of the collector. The points at which the two wires come in contact with the germanium plate are placed very close to each other (not further apart than a few dozens of microns). When point-contact transistors are manufactured, they are subjected to the forming process and then zones of opposite types of conductivity are created inside the germanium, near the points where the basic metal comes into contact with the wires. Thus, pn boundaries exist in a point-contact transistor, just as they do in a junction transistor. The internal capacitance of point-contact transistors is very low and, because of this, such devices can operate on frequencies of several dozen megacycles.

However, as has already been pointed out, the small area of contact in transistors of this type greatly limits their power-handl-

ing ability and makes it possible to use these transistors only in weak-signal circuits.

In contrast to junction transistors, point-contact transistors usually have values of α higher than unity. However, a point-contact transistor has a lower power-amplification factor than a junction transistor, is noted for a higher level of internal noise, and is not quite as robust as the other type of transistor. Amplifier stages with point-contact transistors operate stably only when they use earthed-base circuits. Parasitic oscillation is likely to take place if the circuits chosen are of the earth-emitter or earth-collector variety, and the suppression of such oscillations is not always an easy matter.

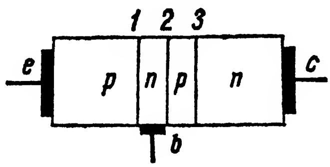


Fig. 309. Junction transistor type *pnpn*, in which α is greater than unity

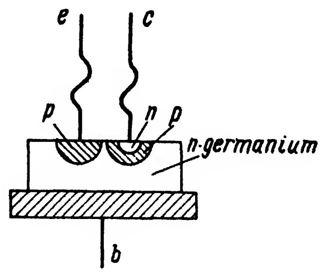


Fig. 310. The structure of a point-contact transistor

For a number of years there has been no satisfactory theory explaining the reason why the value of α in point-contact transistors should be as high as several units in some cases. However, further investigations have brought about a development of junction transistors whose current-amplification factor was made just as high as that of point-contact devices. This was attained by incorporating two thin middle layers with different types of conductivity into the junction transistor (Fig. 309). Such transistors were named *pnpn* transistors. The devices—so far produced only on an experimental scale—possess three *pn* boundaries and act in the following manner.

The holes from the emitter, having passed through the base and collector boundary (*2*) into the thin layer with *p*-type conductivity, are accumulated in the said layer and set up a positive space charge to the left of the *pn* boundary (*3*). This charge acts upon the electrons, an excess of which is present in the *n* zone to the right of boundary *3*; the charge helps the electrons to move through the thin layer *p* located between boundaries *3* and *2*. This transition carried out, the electrons then easily reach the base, because they are acted upon by a strong accelerating field as they pass boundary *2*. In this way, it takes only a comparatively small change of the emitter current to obtain a much greater increase of the collector current. Such an increase is attributed to the flow of electrons through boundary *3*.

As a result, values of α in excess of 1 are obtained. It is the presence of the third boundary in point-contact transistors that accounts for their inherently high values of α. This third *pn* boundary is located near the collector (Fig. 310) and is created as the transistor goes through the formation process during its manufacture.

134. TRANSISTOR CHARACTERISTICS

The properties of various transistors are reflected by their static characteristics. We have already studied the static characteristics of electron valves and know that in the case of a triode valve it is sufficient to have one family of grid or anode characteristics, when the valve draws no grid current. Each one of such families shows the mutual relation between the following three values: anode current, grid voltage and anode voltage. In transistor circuits, we have to deal with four factors which are also mutually related to each other: emitter current and voltage (I_e, U_e) and collector current and voltage (I_c, U_c). Such relation cannot be covered by a single family of characteristics, and two families have to be used for the purpose. For instance, it is possible to study *a family of emitter characteristics* together with *a family of collector characteristics*. When dealing with an earthed-base circuit, incorporating a transistor, these characteristics may be termed, respectively, the *input* and *output characteristics*.

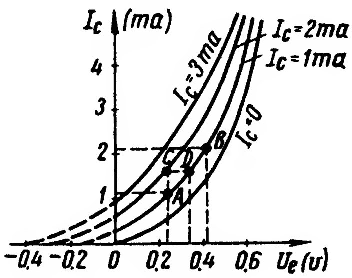


Fig. 311. Approximate shapes of emitter characteristics

Emitter characteristics show the dependence of emitter current I_e upon emitter voltage U_e at some constant value of collector current. Approximate shapes of such characteristics are shown in Fig. 311. As could be expected, these characteristics resemble the volt-ampere characteristics of a semiconductor diode on direct current. As U increases, current I increases, too. Besides, greater values of collector current I_c correspond to greater values of emitter current I_e.

These characteristics are noted for the following peculiarity. Only one curve, corresponding to collector current $I_c = 0$, is seen passing through the zero point of the coordinates. If $I_c = 0$ and $U_e = 0$, then, as can be expected, I_e is also equal to zero. Characteristics corresponding to the cases when $I_c > 0$ pass higher and, as far as they are concerned, the emitter current is not equal to zero even when $U_e = 0$. This phenomenon is most conveniently understood by studying an equivalent of a transistor operating in a

d.c. circuit (Fig. 312). Here, resistances R_e and R_c respectively represent the resistance of the emitter and collector boundaries, while R_b stands for the resistance of the base.

When $U_e = E_1 = 0$ and $U_c = E_2 > 0$, a certain amount of current, caused by the collector voltage, would flow through the collector circuit. And a part of this current will also branch off into the emitter circuit. For instance, the characteristic plotted for the case when $I_c = 3$ ma shows that, when $U_e = 0$, one milliampere of the collector current will branch off into the emitter circuit, the remaining two milliamperes flowing through the resistance of the base. In this case, current I_e can be made equal to zero if negative voltage is applied to the emitter (the whole of the current will then flow through R_b). This is confirmed by the dotted tails of the characteristics shown in Fig. 311.

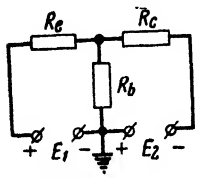


Fig. 312. An equivalent diagram of a transistor for operation in a d.c. circuit

Sometimes, the emitter characteristics are depicted by the mutual substitution of the values given on the X and Y axes, which, of course, does not change the character of the curves. It should be noted that emitter characteristics correspond to very small values of emitter voltage U_e (few fractions of a volt), while the emitter and collector currents are usually given in milliamperes, in case of low-power transistors.

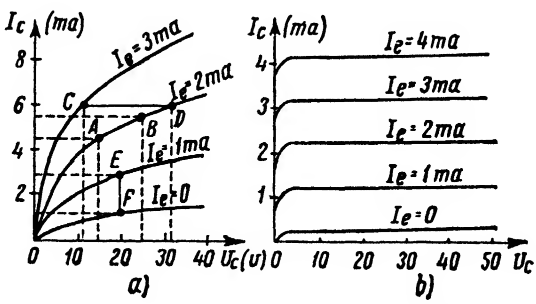


Fig. 313. Approximate collector characteristics of a point-contact transistor (*a*) and a junction transistor (*b*)

Fig. 313 gives examples of collector characteristics of a point-contact and a junction transistor. These characteristics represent the dependence of collector current I_c upon collector voltage U_c at various constant values of emitter current I_e. As may be seen from the curves, the collector characteristics of a point-contact transistor (Fig. 313*a*) possess considerable curved sections and rise

at a great slope. These characteristics resemble the anode characteristics of a positively biased triode valve.

On the other hand, the collector characteristics of a junction transistor (Fig. 313*b*) possess only short initial curved sections, after which the characteristics level out and become nearly horizontal straight lines. These characteristics resemble the anode characteristics of a pentode valve. However, in the case of a pentode, the characteristics look like a diverging bunch of curves, the distance between successive curves varying as the grid voltage is changed in steps of the same value (Fig. 89). The greater the negative voltage at the grid of a pentode, the closer to each other will be the characteristics of the valve. It is because of this, that pentodes introduce considerable distortion when made to amplify high-amplitude signals. Note that the junction transistor is free from this shortcoming; the transistor characteristics are better than the characteristics of a pentode, because they are nearly parallel and pass at identical distances from each other. Therefore, a junction transistor is a better amplifier of high-level oscillations than a pentode valve; unlike the latter, the transistor not only introduces practically no distortion into the signal being amplified, but it also offers a higher efficiency, both of which advantages are impossible to obtain with a pentode.

As a matter of fact, the collector characteristics of a transistor are the same as the reverse-current volt-ampere characteristics of a diode (Fig. 292). Because of this, some designers consider U_c and I_c as negative values, plotting them along the left-hand and the lower parts of the coordinate axes. In some cases, the positions of U_c and I_c are interchanged on the axes.

It may be seen from the characteristics that, as U_c increases, the current at first sharply rises, after which the rise becomes slower (stopping almost altogether in case of junction transistors). Thus, a peculiar state of saturation sets in. This is explained by the fact that the reverse current is created by the secondary carriers. Because the number of such carriers in a semiconductor is rather small, they are all used up even when the values of U_c are comparatively low. If the collector current is to be increased further, it becomes necessary to increase the emitter current. Then the base will be reached by additional secondary carriers from the emitter boundary (such secondary carriers being the holes in the case of a *pnp* transistor). The characteristics serve as a pictorial representation of collector-current increase, this current rising in accordance with the increase of I_e. Higher values of I_e cause upward displacement of the curves. In case of collector characteristics, voltage U_c is generally of the order of some dozens of volts for normal values of current I_c.

Besides the collector and emitter characteristics, other types of characteristics are also sometimes given. Among these, *amplifi-*

cation characteristics, otherwise known as *transfer characteristics*, show the dependence of voltage U_c upon current I_e for various constant values of I_c. The name "amplification characteristics" is derived from the fact that, as a result of operation of an amplifier stage, the value of U_c varies under the influence of I_e changes. Another type of characteristics, usually studied together with the amplification characteristics, are known as *feedback characteristics*. These characteristics represent the dependence of voltage U_e upon current I_c for various constant values of current I_e. It is learnt from these characteristics how collector-current changes cause variations of emitter voltage. This effect is attributed to the considerably close feedback coupling existing inside a transistor, which is not difficult to understand if we consider that collector current I_c unavoidably flows through the emitter circuit, the changes of this current influencing the values of voltage U_e.

Apart from the above-named types of characteristics, still other characteristics can also be employed. Generally speaking, opinions still vary as to which characteristics are the most convenient in the calculation of transistor circuitry. Various designers, being guided by one reason or another, offer to make use of different characteristics, and there is no unanimity in this respect. All this is quite natural at the present stage of development of the transistor art. The final theory of transistors is still in the making and opinions will differ for some time to come, until the exact nature of various phenomena taking place in the semiconductor devices has been thoroughly evaluated and understood.

135. TRANSISTOR PARAMETERS AND EQUIVALENT CIRCUITS

There are various systems of static transistor parameters offered by different designers for the calculation of transistor circuitry. From these systems we shall choose, first of all, the system of the so-called *Z-parameters*, where the parameters are represented by certain resistances. These resistances are not exactly pure ohmic resistances. However, for the sake of simplicity, the reactive components are usually disregarded when these resistances are dealt with, which is particularly justifiable on low frequencies.

Taking the case of an earthed-base transistor circuit, and employing figures 1 and 2 as respective indices to designate the emitter and the collector, the definitions of the Z-parameters can be formulated as follows:

No-load input resistance, r_{11}, is the relation between emitter voltage change ΔU_1 to emitter current change ΔI_1, the latter change caused by the ΔU_1 change, while the collector current is kept at a steady value,

or: $$r_{11} = \frac{\Delta U_1}{\Delta I_1}; \qquad I_2 = \text{const.}$$

In other words, r_{11} is the input resistance for the a.c. component of the emitter current, when the collector current is not changing.

Designating the a.c. components by small letters, the above expression may be rewritten as follows:

$$r_{11} = \frac{u_1}{i_1}, \text{ when } i_2 = 0.$$

In the definition of resistance r_{11}, the words "no-load" signify that $i_2 = 0$, i.e., the condition as if the output circuit were open, as far as the alternating current is concerned.

The value of r_{11} does not exceed several hundred ohms. The exact value of this resistance can be easily determined from the emitter characteristics, as follows. Take the section of some characteristic (for instance, section AB in Fig. 311). Note the ΔU_e and ΔI_e values corresponding to this section. Dividing these values one by another will give the exact value of r_{11} for any desired point of the characteristic. It should be noted that the values of r_{11} will not remain the same for different sections of the characteristic; these values will vary in accordance with changes in the operating conditions. This holds true also when dealing with all the other parameters of a transistor.

Reverse no-load output resistance is the second important parameter of a transistor. This parameter is designated by r_{22} and is found from the following formula:

$$r_{22} = \frac{\Delta U_2}{\Delta I_2}, \text{ when } I_1 = \text{const.}$$

or

$$r_{22} = \frac{u_2}{i_2}, \text{ when } i_1 = 0.$$

This expression defines the output resistance for the a.c. component of collector current when no changes take place in the emitter current.

In the case of point-contact transistor, this output resistance may range from some kilohms to several dozens of kilohms, while in the case of junction transistors the upper limit is several hundred kilohms or even some megohms. The output resistance can be found from the collector characteristics by the same method which was used above to determine the value r_{11} (as an example, use points A and B in Fig. 313 when determining the value of r_{22}). The value of r_{22} depends upon the slope of the collector characteristics. The steeper the slopes, the greater the value of r_{22}, referring to the way the collector characteristics are represented in Fig. 313.

Amplification resistance, r_{21}, is the next important parameter of a transistor and is found from the following formula:

$$r_{21} = \frac{\Delta U_2}{\Delta I_1}, \text{ when } I_2 = \text{const.}$$

or

$$r_{21} = \frac{u_2}{i_1}, \text{ when } i_2 = 0.$$

The value of r_{21} is of the same order as the value of r_{22} and characterises the building up of amplified alternating voltage in the collector circuit as the emitter current is varied. The actual value of r_{21} may be found in the following way. Keeping I_c at the same constant value, select two points on two different collector characteristics (for instance, points C and D in Fig. 313). This done, divide the values ΔU_c and ΔI_e by each other, these values corresponding to the given section of the characteristics. This will give the actual value of r_{21}.

Feedback resistance, r_{12}, is the last parameter of the group of transistor parameters we are here discussing. This parameter is determined from the following formula:

$$r_{12} = \frac{\Delta U_1}{\Delta I_2}, \text{ when } I_1 = \text{const.}$$

or

$$r_{12} = \frac{u_1}{i_2}, \text{ when } i_1 = 0.$$

We have already noted above that the internal feedback is typical of all transistors. The a.c. component of collector current, passing through various

elements of the emitter circuit, builds up a certain a.c. voltage in the given circuit. It is this feedback that is characterised by resistance r_{12}. The value of r_{12} is of about the same order as the value of r_{11} and can be found from the emitter characteristics. To do this, take the ratio of increments ΔU_e and ΔI_c, these increments corresponding to the section between two points on two different characteristics (for instance, points C and D in Fig. 311). The value of emitter current I_e should be kept steady during this measurement.

The four transistor parameters discussed above should be supplemented with current amplification (or current gain) factor α, which we have studied previously. Using the designations given above, the following formula may be written for α:

$$\alpha = \frac{\dot{i}_2}{\dot{i}_1}, \text{ when } u_2 = 0.$$

The value of α can be easily found from the collector characteristics. To do this, take two different curves and select on them the points standing for the same value of Uc (for instance points E and F in Fig. 313). The ratio of increments ΔI_c and ΔI_e, taken between these points, is equal to α.

Using the aforesaid parameters, the relation between a.c. components of currents and voltages in the circuit of a transistor can be expressed by the following two equations:

$$u_1 = i_1 r_{11} + i_2 r_{12},$$

$$u_2 = i_1 r_{21} + i_2 r_{22}.$$

The first of these two equations indicates that the alternating voltage in the emitter circuit is made up of the voltage drop produced by current i_1 across input resistance r_{11}, and the feedback voltage built up by current i_2 flowing through resistance r_{12}.

As follows from the second of these two equations, the alternating voltage of the collector is made up of the voltage drop produced by current i_2 across output resistance r_{22}, and of another voltage drop built up by current i_1 flowing through resistance r_{21}.

In accordance with these two equations, it becomes possible—as far as the a.c. components are concerned—to represent the transistor by the equivalent circuit shown in Fig. 314 and actually consisting of two separate circuits.

Studying this equivalent circuit, we assume the following. The source generating the oscillations to be amplified is connected to the left-hand (input) terminals, while the load resistance is connected across the right-hand (output) terminals. No power supplies are shown, as we deal with the distribution of alternating current only in this theoretical circuit (we simply assume that the a.c. resistance of such power supplies is sufficiently low to be neglected).

The input part of the circuit contains the equivalent of a generator developing a voltage of $i_2 r_{12}$. This generator acts to simulate the action of the internal feedback inside the transistor. Another equivalent generator is connected in the output part of the circuit to show that amplified voltage appears in this part. The voltage developed by this generator is given as $i_1 r_{21}$. It should not be considered that r_{12} and r_{21} stand for the internal resistance of the two generators. These parameters simply play the role of proportioning factors between respective currents and voltages. For instance, multiplying current i_2 by resistance r_{12} we obtain the feedback voltage value, this voltage being proportional to i_2. It is, of course, clear that such a proportioning factor must have the dimensions of resistance, while—as has already been stipulated—we consider the internal resistance of our equivalent generators as zero. In other words, in the theoretical circuit under discussion we shall have to assume that voltages developed by the generators are their electromotive forces.

Although in the circuit given in Fig. 314 the input and the output parts of the circuit are shown quite separately, an interdepedence exists between them, which is taken into account when discussing the operation of the equivalent generators.

Speaking of equivalent circuits in general, the following observations may be made. The analytical study of transistors, depicted in the shape of equivalent circuits, has brought about many different ways of representing such circuits. There is, however, still much debating as to which of these several ways are more convenient and logical. In particular, the circuit which we have already noted (Fig. 314) has a shortcoming in that it does not depict the actual action of the electric circuit of the transistor. Because of this, it is generally considered a better practice to employ, in its stead, the equivalent circuit given in Fig. 315. In this circuit, r_e, r_c and r_b stand, respectively, for the resistances of emitter and collector boundaries and the base (as far as the alternating current is concerned). The circuit of Fig. 315 resembles the circuit of Fig. 312. However, these two circuits should not be confused, because the one of Fig. 312 is not suitable for

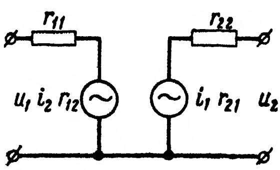


Fig. 314. The equivalent a.c. circuit of a transistor

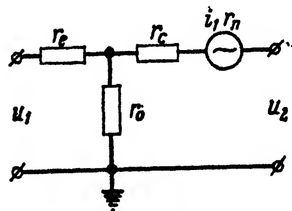


Fig. 315. The equivalent circuit of a transistor with an earthed base

the discussion of a.c. circuits. It is unsuitable because it comprises resistors R_e, R_c and R_b, *with reference to d.c. only*, while resistors r_e, r_c and r_b, in which we are at present interested, represent resistance values for alternating current. Besides all this, the circuit of Fig. 312 does not indicate that amplified a.c. voltage is present at the collector. If a source of oscillatory voltage is connected to the input part of such a circuit, there will appear at the output, not an amplified voltage, but a voltage smaller than the one applied to the input (in this case, the loss of voltage will take place in resistances R_e and R_c).

Having noted the shortcomings of various equivalent circuits (including the circuits of Fig. 312 and Fig. 314), let us now come to the problem on hand, employing the circuit of Fig. 315 as the most suitable reference for our discussion. In this circuit, the amplified voltage at the output is obtained from a certain equivalent generator connected into the collector circuit. The voltage (or rather the e.m.f.) developed by this generator is proportional to emitter current i_1. The proportioning factor between these two values, which necessarily possesses the dimensions of resistance, is denoted as r_n and may be called the *transfer resistance*, although many designers of transistor circuitry call it by other names and denote it differently. It should be kept in mind that r_n is not the internal resistance of the equivalent generator (the value of r_n is equal to the value of αr_c).

At this point it is well to recall the electron valve amplifier stages and their functions. As we know, in the equivalent circuit related to the alternating anode-current of such a stage (Fig. 132*b*), the e.m.f. of the equivalent generator is obtained by multiplying the value of a.c. grid voltage by the amplification factor of the valve. In this case, the alternating e.m.f. acting in the anode circuit is proportional to the grid voltage. Because of this, when dealing with valve amplifier circuits, the proportioning factor is only a relative value having no dimensions.

If we define alternating anode current as a function of a.c. voltage at the grid of a valve, then mutual conductance of the valve will be the proportioning factor. This factor, though it has the dimensions of conductivity, does not represent the actual conductivity of any particular part of the circuit.

When dealing with a transistor, a similar role is played by transfer resistance r_n. Only, in this case, referring to Fig. 315, the internal resistance of the generator (developing an e.m.f. of $i_1 r_n$) is equal to zero.

Resorting again to the equivalent circuit of Fig. 315, we introduce new parameters in the shape of four resistances r_e, r_c, r_b and r_n. These parameters are in very simple relations with the parameters we had previously determined and denoted as r_{11}, r_{22}, r_{12} and r_{21}. As we may recall, the equivalent circuit given in Fig. 314 pertains to the last four parameters.

In reality, both equivalent circuits represent the same transistor and, hence, should provide similar results. Comparing the two circuits with each other, we can write the following equations:

$$r_{11} = r_c + r_b; \qquad r_{22} = r_c + r_b;$$

$$r_{12} = r_b; \qquad r_{21} = r_n + r_b.$$

Hence:

$$r_e = r_{11} - r_{12}; \qquad r_c = r_{22} - r_{12};$$

$$r_b = r_{12}; \qquad r_n = r_{21} - r_{12}.$$

It should be noted that the given equations hold true only for the earthed-base transistor circuits. When earthed-emitter or earthed-collector amplifiers are employed,other types of equivalent circuits are obtained with different values of r_{11}, r_{22}, r_{12} and r_{21}. Because of this, r_{11}, r_{22}, r_{12} and r_{21} are only secondary parameters.

As far as resistances r_e, r_c, r_b and r_n are concerned, these resistances pertain to the transistor itself and, hence, can be considered as primary parameters. Below are listed the typical values of these primary parameters.

Emitter resistance r_e ranges between some dozen ohms in junction transistors and several hundreds of ohms in point-contact transistors.

Base resistance r_b is of the order of several hundred ohms for both types of transistors, the higher values predominating in junction transistors.

The resistance values of r_c and r_n are considerably higher in both types of transistors, running from several kilohms to several dozen kilohms, in case of point-contact transistors, to hundreds of kilohms and even some megohms in junction transistors.

In the above discussion we have been considering the so-called Z-parameters of junction and point-contact transistors. However, the resistance comparison method we have employed is not the sole method of evaluating transistors. Many designers of transistor circuitry recommend the use of conductivity comparison method, which is known as the Y-parameter method. When dealing with junction transistors, the Y-parameter method offers certain conveniences not provided by the Z-parameter method and facilitates the design of the circuitry.

In recent years still another method of transistor evaluation has come to the force. This method is known as the method of *hybrid parameters*, in which the parameters are denoted by letter h, to which an appropriate index is attached. The name "hybrid" is used in this case, because among these parameters appear two non-dimensional values: one resistance value, and the other a conductivity value.

These parameters have proven to be convenient in use both in the study of junction and of point-contact transistor. Two of these parameters are determined by short-circuiting the output circuit for a.c., i.e., when $u_2 = 0$.

1. *Short-circuit input resistance* h_{11}.

$$h_{11} = \frac{u_1}{i_1}, \text{ when } u_2 = 0,$$

h_{11} is approximately equal to emitter resistance r_e.

2. *Current-gain factor*, h_{21}.

$$h_{21} = \frac{i_2}{i_1}, \text{ when } u_2 = 0.$$

This parameter represents the α parameter which we have already discussed. The other two parameters are determined when the input circuit is open, as far as a.c. is concerned, i.e., when $i_1 = 0$.

3. *Feedback factor*, h_{12}.

$$h_{12} = \frac{u_1}{u_2}, \quad \text{when } i_1 = 0.$$

The feedback factor is approximately equal to the ratio of r_b and r_c, i.e., $h_{12} \approx \frac{r_b}{r_c}$.

4. *Reverse no-load output conductivity*, h_{22}.

$$h_{22} = \frac{i_2}{u_2}, \text{ when } i_1 = 0,$$

h_{22} is approximately equal to the reciprocal value of collector resistance, i.e., $h_{22} \approx \frac{1}{r_c}$.

135. BASIC CIRCUITS OF TRANSISTORISED AMPLIFIERS

As noted at the outset of our discussion of transistors, there may be three general versions of transistorised circuits: the earthed-base circuit, the earthed-collector circuit and the earthed-emitter circuit. In the analysis just given above we have treated in some detail the first version, i.e., the earthed-base arrangement. This has given us a preliminary understanding of transistorised circuits in general. We shall now summarise the most prominent features of all the three versions. Comparing these versions to each other, we shall see our way clear to practical application of each particular version to any specific problem on hand.

The earthed-base amplifier. An amplifier of this type, shown in Fig. 316*a*, is noted for low input resistance r_i and high output resistance r_o. Here, the value of r_i ranges within a few dozen ohms

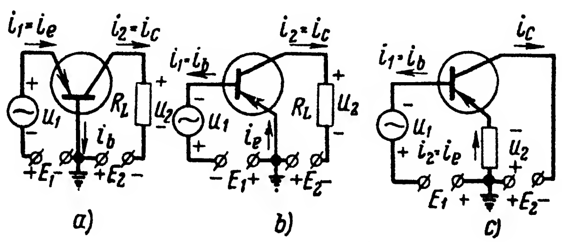


Fig. 316. Circuits of amplifier stages with: earthed base *(a)*; earthed emitter *(b)*; and earthed collector *(c)*

for transistors of either type, while the values of r_o may be anything within several kilohms for point-contact transistors to several hundreds of kilohms for junction transistors. Current gain coefficient c_i is somewhat smaller than unity for junction transistors, and

greater than unity for point-contact devices. Voltage-gain coefficient c_u of the amplifier stage reaches several tens (several hundreds in the case of junction transistors), while power gain coefficient c_p may be as high as one hundred for point-contact transistors or several hundreds in the case of junction transistors. The phase of oscillations is not inverted in amplifiers of this type, and the earthed-base circuit has the best stability of all the three circuit versions. This circuit is the only one in which point-contact transistors will operate stably (junction transistors provide stable operation in any one of the three circuit versions).

Generally speaking, the instability of operation of some transistor circuits is attributed to the presence of the internal feedback, which, as we have already noted, exists in all transistors. The feedback effect arises owing to the fact that one of the inner resistances of the transistor (for instance, resistance r_b in the earthed-base circuit) is connected simultaneously into the input and output circuits, acting like a feedback resistor. Alternating current i_2 of the output circuit flows through this resistor and builds up a feedback voltage across it, this voltage influencing the input circuit. If the feedback happens to be positive and sufficient, the stage will excite itself and will act as an oscillator and not as an amplifier.

The feedback voltage increases with the increase of current i_2 and with the increase of the feedback resistance. In the earthed-base circuit, the feedback resistance is represented by base resistance r_b. The value of this resistance is low. Therefore, the feedback is weak and the circuit, as a rule, operates stably. Still, in some cases, when point-contact transistors are used in such a circuit, the stage will oscillate because the value of α is greater than unity in such transistors and, accordingly, their current i_2 is considerable and exceeds current i_1. It should be noted that i_2 increases when load resistance R_L is lowered and reaches its maximum value when $R_L = 0$. Thus, in the case of a short circuit in the output of the stage, the feedback is intensified and the stage is most apt to break into oscillation.

Modern point-contact transistors are noted for sufficiently low value of r_b. Because of this, they, as a rule, provide stable operation in the earthed-base circuit, even when the value of R_L is small. The stability of junction transistors is assured by the fact that in these transistors $\alpha < 1$ and $i_2 < i_1$, which makes the feedback too weak to cause self-excitation of the stage. The stability of the earthed-base amplifier stage will be impaired if an additional resistance is connected into the base lead. Such resistance, together with r_b, will play the role of the feedback resistor. The stage will break into oscillation at a certain value of such resistor. On the other hand, the stability of the circuit will be further improved if higher values of resistances are connected into the collector and emitter circuits.

The earthed-emitter amplifier. This amplifier (Fig. 316*b*) is analogous to the earthed-cathode valve amplifier stage. The input resistance of the earthed-emitter amplifier is considerably greater than that of the earthed-base amplifier. In the stage here described, the value of r_i can reach dozens of kilohms in the case of point-contact transistors, and from several hundred ohms to about 1.5 kilohms in the case of junction transistors. The output resistance is expressed in several dozens of kilohms, the higher values being encountered in junction transistors. In the circuit being discussed, current-gain coefficient c_i can reach several dozens in the case of junction transistors. This follows from the fact that in this amplifier circuit output current i_2, equal to collector current i_c, is considerably greater than base current i_b (the latter being the input current i_1), while, as we know, c_i is the ratio $i_2 : i_1$.

Taking into consideration that: $i_b = i_e - i_c$, and also that current gain coefficient in the earthed-base circuit is given by: $\alpha = \frac{i_c}{i_e}$, the value of c_1 may be expressed as follows:

$$c_i = \frac{i_2}{i_1} = \frac{i_c}{i_b} = \frac{i_c}{i_e - i_c} = \frac{i_c}{i_c\left(1 - \frac{i_c}{i_e}\right)} = \frac{\alpha}{1 - \alpha}.$$

If, for example, $\alpha = 0.95$, then

$$c_i = \frac{0.95}{1 - 0.95} = \frac{0.95}{0.05} = 19.$$

The earthed-emitter circuit has a voltage gain coefficient of several hundreds, while its power gain coefficient reaches several thousands. High gain is the main advantage of this circuit.

Fig. 316*b* shows the polarity of a negative half-wave of the voltage developed by the source of oscillations. This is the polarity at which amplification of currents i_e and i_1 takes place. During this process, current i_2 increases. This produces a positive half-wave of a.c. voltage across load resistor R_L. (The arrows in the circuit of Fig. 316*b* are related to the a.c. components of the currents flowing in the circuit.) Thus, the phase of the alternating voltage is reversed in the earthed-emitter stage. The possibility of feeding such a stage from a single supply source (E_2) is an additional advantage of the earthed-emitter circuit.

Owing to their advantages, earthed-emitter amplifiers have become quite popular. It should be noted, however, that the frequency response of these amplifiers is worse than that of the earthed-base circuits: as the frequency is increased, the gain of the earthed-emitter amplifier drops off faster than it does in the previously discussed amplifier. Besides, only junction transistors may be employed in the earthed-emitter amplifiers; any attempts at using point-contact transistors in such amplifiers will lead to parasitic self-oscillations.

The earthed-collector amplifier. The circuit of such an amplifier (Fig. 316*c*) is analogous to the circuit of the valve amplifier in which the anode is earthed. As we know, such valve amplifiers are called cathode followers. It, therefore, would be logical to name the transistorised amplifier here described as *emitter follower.* This amplifier operates quite stably only with junction transistors. The main advantages of the "emitter follower" amplifier are its very high input resistance (hundreds of kilohms), and high current gain. On the other hand, the output resistance of this amplifier is low and does not exceed some tens of ohms for junction transistors and hundreds of ohms for point-contact transistors.

In the earthed-collector amplifier, the voltage phase is not inverted during the amplification process, while the voltage-gain coefficient of the amplifier is somewhat less than unity. Current gain coefficient c_i for junction transistors can reach several dozens. The formula for c_i is easily obtained if we take into consideration that in the given circuit the following relations exist: $i_2 = i_e$; $i_1 = i_b$; $i_b = i_e - i_c$, i.e.,

$$c_i = \frac{i_2}{i_1} = \frac{i_e}{i_b} = \frac{i_e}{i_e - i_c} = \frac{i_e}{i_e\left(1 - \frac{i_c}{i_e}\right)} = \frac{1}{1-\alpha}.$$

For instance, if $\alpha = 0.95$, $c_i = 20$.

The earthed-collector amplifier provides a relatively low power gain and is not employed as much as the earthed-base and the earthed-emitter amplifiers.

137. SOVIET TRANSISTORS

The Soviet Union manufactures various types of point-contact and junction transistors.

The design of the earlier point-contact transistors is shown in Fig. 317*a*. The devices are enclosed in metal bodies, which simultaneously serve as base terminals. The emitter and collector terminals are pin-shaped. Transistors shown in Fig. 317*a* come in two series, known as series C1 and series C2.

There are six type varieties in the C1 series: the transistors of various types ranging from C1A to C1E designed to amplify oscillations with the upper frequency limit from 0.5 to 10 mc. The transistors of this series operate at collector voltages from 20 to 40 volts. Their current gain coefficient α is between 1.2 and 1.5 on low frequencies and drops to 1.0-1.2 values at the highest frequency of operation. An amplifier stage employing such transistors can provide voltage amplification of 30-50. When the stage is terminated with a 10,000-ohm load resistor, the power gain can be 30-100. Power dissipated at the collector must not exceed 50-100 milliwatts.

The C2 series is comprised of four type varieties, ranging from C2A to C2Г. The upper frequency limit of this series is the same as that of the C1 series (0.5 to 10 mc), but the transistors of the C2 series are primarily intended to act as oscillators. Their collector voltage can range from 10 to 30 volts. The value of α is between 1.6-1.5 on low frequencies and may equal 1.5-1.2 at the highest operating

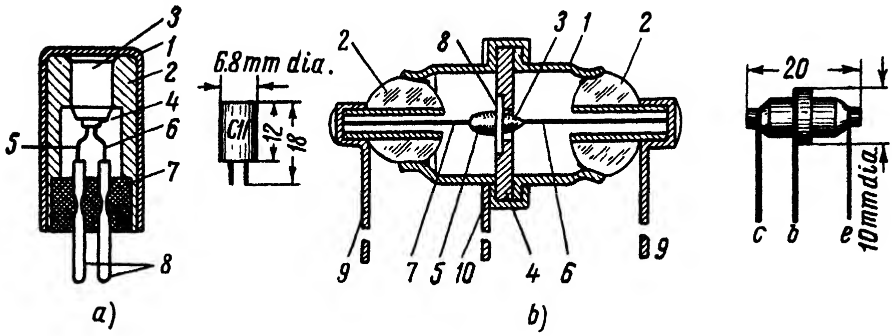


Fig. 317. The design principle and external view of point-contact and junction transistors: *a)* point-contact transistors, series C1 and C2: *1* — cover; *2* — body (base); *3* — crystal holder; *4* — germanium; *5* — emitter; *6* — collector; *7* — insulating bush; *8* — terminals; *b)* junction transistors, series П1 and П2: *1* — body; *2* — glass; *3* — indium (emitter); *4* — crystal holder; *5* — indium (collector); *6* — emitter terminal; *7* — collector terminal; *8* — germanium; *9* — external terminals of emitter and collector; *10* — base terminal

frequency. Maximum dissipation power of the collector of such a transistor is between 50 and 100 milliwatts.

The latest point-contact transistor series, manufactured at present, carry designations of C3 and C4. These transistors are noted for an improved design and are hermetically sealed in metal cases (Fig. 318). The electrical parameters of the C3 series are similar to those of the C1 series, the new series extending in type varieties from C3A to C3E (six varieties). The C4 series, extending from C4A to C4Г, possesses electrical parameters similar to the C2 series.

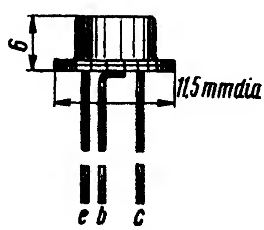


Fig. 318. External view of point-contact transistors C3 and C4 and of junction transistors П6

All types of point-contact transistors can operate within a temperature range of —50° C to +60° C.

The following situation exists in the manufacture of junction transistors.

First, the industry is putting out П1, П5 and П6 junction transistors for the purposes of voltage amplification. The earlier triodes of the П1 series, ranging from П1А to П1Д, are used for the amplification of signals whose frequency does not exceed 100 kc. An extension

of this production line ranges up to type П1Е transistor, whose upper frequency limit is 465 kc, П1Ж transistor—with the limit of 1 mc, an П1И transistor—with the limit of 1.6 mc. The П series transistors are enclosed in metal cartridges with glass insulators (Fig. 317*b*). The germanium plate in these transistors is shaped as a washer pressed into the cartridge and soldered to the holder. Therefore, the body terminal is also the base terminal. The indium impurity is infused into the germanium from both sides, emitter and collector terminals being soldered to the indium. These terminals pass through the glass insulators mentioned above. The П series transistors operate with collector voltages from 10 to 20 volts. The value of α ranges from 0.93 to 0.97 on low frequencies and may equal 0.7 at the highest operating frequency. When a transistor of this series is employed in the earthed-emitter circuit on a frequency of 1 kc, and when the load resistance is equal to 30.000 ohms, power gains as high as 1.000-5.000 can be obtained. The maximum dissipation power of the collector is 50 milliwatts.

The latest versions of the П series are type varieties extending from П5А to П5Д. This is the so-called П5 series (Fig. 319). The transistors of this series are enclosed in midget glass envelopes and their collector terminal is marked with a red dot. The collector voltage of these transistors is between 2 and 10 volts. The upper frequency limit at which these transistors are still capable of providing satisfactory amplification varies from 100 to 500 kc, depending upon the type of transistor. The value of α at the upper frequency limit varies between 0.93 and 0.995. The collector dissipation should not exceed 25 milliwatts.

Five types of transistors, ranging from П6А to П6Д are also new improved-design versions. These units are enclosed in sealed metallical bodies (just like the transistors shown in Fig. 318) and are intended for amplification service at a frequency limit ranging from 0.5 to 2.5 mc, depending upon the transistor type. The collectors of these transistors operate at voltages from 5 to 30 v. The maximum collector dissipation is 150 milliwatts. When used in the earthed-emitter circuit, and when the load resistance is equal to 30,000 ohms. power gains as high as 3,000-10,000 are obtained with these transistors. Their α value is equal to 0.92.

Specially designed junction transistors, known as types П2, П3, П4 an П7, are intended for amplification of low frequencies only. Earlier types of these transistors, known as П2А and П2Б, are shown in Fig. 317*b*. The collector voltage of П2А transistor is between 50 and 100 v, that of П2Б transistor—between 25 and 50 v. The value of α is at least 0.9. Maximum power that can be dissipated at the collector is 250 milliwatts. When the transistor of this type is used in the earthed-base circuit, output power of 100 milliwatts may be obtained, at less than 15% non-linear distortion. The power gain under these conditions is of the order of 50.

Transistor types П3А, П3Б and П3В are noted for higher power ratings. Their external view and dimensions are given in Fig. 320. They operate at collector voltages between 12 and 50 volts. When used in the earthed-emitter circuit, they are capable of power output up to at least one watt, the non-linear distortion factor not exceeding 15%. The power gain factor of these transistors, when operated under conditions just described, is as follows: 50—for П3А device, 100—for П3Б device, and 300—for П3В device.

New high-power transistors come in five types, classified as types П4А, П4Б, П4В, П4Г, П4Д. When one of these improved transistors is operated in a single-ended earthed-emitter class A

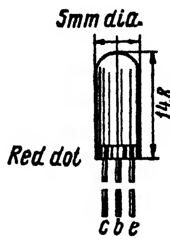


Fig. 319. External view of junction transistors П5

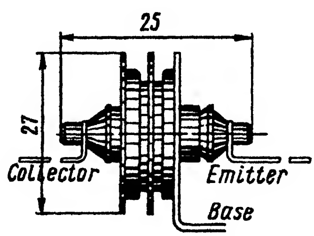


Fig. 320. External view of junction transistors П3

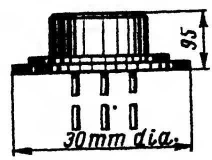


Fig. 321. External view of power junction transistors П4

circuit, it is capable of developing an output power of at least 10 watts. Two such transistors, employed in a push-pull class B circuit, give an output of at least 30 watts. These transistors operate at collector voltages from 26 to 60 volts. Constructionally, they are represented by welded, flanged-base metal containers (Fig. 321). Electrode terminals are passed through glass insulators installed in the base.

Type П7 transistor is enclosed in a midget glass envelope (like the transistor shown in Fig. 319), but the envelope is slightly shorter than the one used by transistors П5. When arranged to operate in a push-pull circuit, two type П7 transistors give an output of at least 0.2 watts. The cooling of these transistors is improved by the metal-wrap around their envelopes. These devices operate at collector voltages ranging from 4.5 to 13 v. Their α value ranges between 0.97 and 0.995.

As a rule, junction transistors can operate within the ambient temperature limits extending from —60 C° to +50°C. All Soviet-made transistors belong to the *pnp* category. It should be noted that they all can operate at lower collector voltages than those given above. The power output of the devices will, of course, be reduced under such operating conditions.

188. TRANSISTORISED RADIO CIRCUITS

General

Transistors are successfully used in place of electron valves in practically all types of radio equipment, with the exception of the equipment designed to operate on extremely high frequencies. Thus, transistors have found a wide application in low-frequency amplifiers, radio receivers, oscillators, radio transmitters, TV receivers, measuring instruments, various pulse-circuits, electronic computing machines, and in many other types of apparatus.

Radio and electronic circuits which were previously employing electron valves but at present use transistors in place of the valves are called *transistorised circuits*. A transistorised circuit consumes a much smaller amount of electric energy than does its valve counterpart, and, besides, is much more compact. If, for example, the minimum energy required for feeding an electron valve is equal to 0.1 watt and more, the minimum energy consumed by an equivalent transistorised circuit amounts only to approximately one microwatt. This gives a 100,000-time energy saving, which is quite unprecedented in the history of radio equipment design practice. Replacement of thousands of electron valves by transistors in such a complex piece of equipment as an electronic computing machine cuts down the energy consumption from 3 kilowatts to a mere 100 watts, i.e., by 30 times.

Transistorised circuits are particularly favoured by designers of portable receivers and transmitters, where every watt of battery power is at a premium. In such applications, the transistorised equipment may be frequently operated from the ordinary torchlight battery.

Radio manufacturing industry has fully utilised the advantage of the equipment miniaturisation offered by the transistorised circuitry. Special superminiature components (resistors, capacitors, coils, etc) are being produced for the express purpose of their application in transistorised equipment. This has made it possible to design radio stations of almost unbelievably small dimensions and weight. Thus, cases have been recorded when a whole transmitting-and-receiving radio station is contained in the usual telephone handset. Due to its unusually low power consumption, such a station is powered by the energy of the human voice and requires no batteries of any kind. Further development of new radio and electronic equipment will, no doubt, lead to the creation of other spectacular types of devices, making full use of the miniaturisation and energy-economy features offered by transistorised circuitry.

Even by this time, when the art of transistorised circuitry is very young, there exists a great variety of interesting circuits based solely on transistors and employing no electron valves of any kind. It would take a whole book to describe all such circuits and we are, therefore, forced to limit our acquaintance with transistorised equipment to a few representative types of such equipment.

Fig. 322 gives the circuit diagram of a two-stage resistance-coupled amplifier, in which both stages use the earthed-base arrangement. Separate power supplies are employed to feed the emitter and collector circuits. Since the input resistance of an earthed-base stage does not exceed a few hundred ohms, the load resistor of the first stage will have an equally low resistance value. Therefore, voltage gain coefficient c_u of this stage is very low. Because of the low input resistance r_i, the capacitance of coupling capacitors C_1 and C_2 has to be made large (about 10 mfd). When type CIΓ transistors are used in the circuit being described, proper operation of the circuit calls for the following values of feeding voltages and components: $E_1 = 20$ v; $E_2 = 5$ v; $R_1 = R_3 = 4.8$ kilohms; $R_2 = R_4 = 6.8$ kilohms; $C_3 = C_4 = 20$ mfd. Voltage gain coefficients of the two stages are, respectively: $c_1 = 2$; $c_2 = 20$. The overall gain of the amplifier is equal to 40. The low amplification obtainable in the first stage is the disadvantage of the given amplifier circuit.

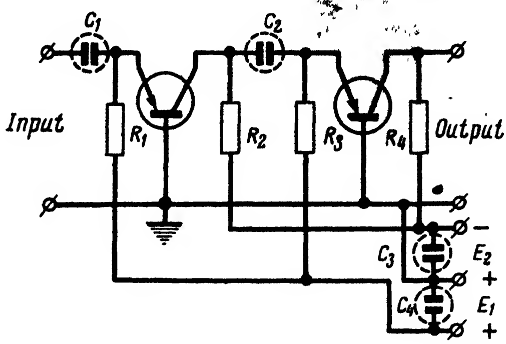


Fig. 322. The circuit diagram of a two-stage low-frequency amplifier

In the aforesaid earthed-base arrengement, transformer coupling gives a better account of itself. This type of coupling, using proper types of step-down transformer, makes it possible to match the low input impedance of the following stage with the high output impedance of the preceding stage. This results in greater power, and voltage, gains than those which the resistance-coupling arrangement can provide. Thanks to the miniature values of currents, the transformers can be made very small for application in such a circuit. The circuit of a two-stage transformer-coupled amplifier is given in Fig. 323. A single power supply E is the peculiarity and advantage of this circuit. Positive bias voltage is automatically fed to the emitters with the help of resistors R_1 and R_2, which are shunted by capacitors. The voltage drop across these resistors is created by the d.c. currents of the two bases.

The circuit diagram of a two-stage earthed-emitter amplifier is shown in Fig. 324. A circuit of this type will provide stable operation only when it employs junction transistors. The first stage of this circuit gives a high voltage amplification, because the input impedance of

an earthed-emitter stage is quite high. The demand for only a single power supply is the great convenience of this circuit. When the circuit uses type П1Г transistors, the following values of components

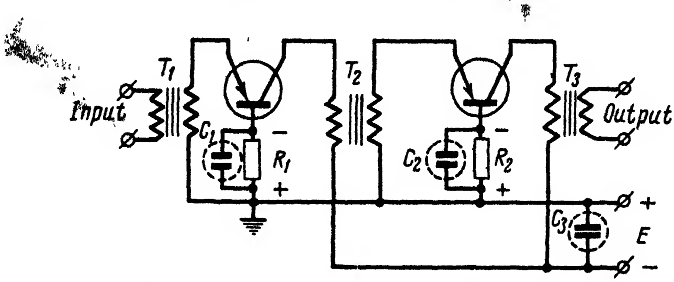


Fig. 323. The circuit diagram of a two-stage transformer-coupled amplifier

and supply voltage must be used: $E=30$ v; $R_1=R_3=220$ kilohms; $R_2=R_4=10$ kilohms; $C_1=C_2=1$ mfd; $C_3=10$ mfd. The overall voltage gain of the amplifier just described is about 400.

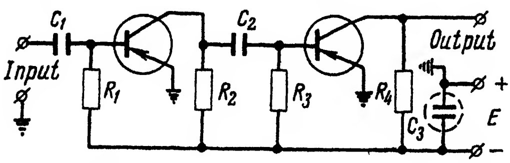


Fig. 324. The circuit diagram of a two-stage earthed-emitter amplifier

As seen from the above examples, low-frequency amplifiers usually employ earthed-emitter or earthed-base circuits. Multi-stage amplifiers sometimes use earthed-collector circuits in individual stages. A stage of this type gives no voltage gain but serves as a matching link. It has a high input impedance, which is always desirable; while the low output impedance of the stage makes it possible to use the earthed-base arrangement in the following stage.

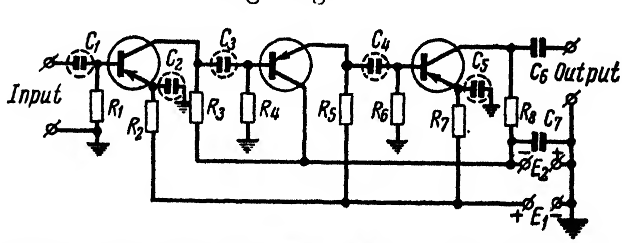


Fig. 325. The circuit diagram of an amplifier with alternate earthed-emitter and earthed-collector stages

An example of such an amplifier is shown in Fig. 325. In this amplifier, two earthed-emitter stages are coupled to each other through an earthed-collector stage. The amplifier employs junction

transistors and gives a voltage gain of about 1,000 while passing a 100-kc band of frequencies. The following specifications are applicable to this circuit: $C_1 = C_2 = C_4 = C_5 = 50$ mfd; $C_3 = 5$ mfd; $C_6 = 2$ mfd; $C_7 = 2$ mfd; $R_1 = R_6 = 10$ kilohms; $R_2 = R_4 = R_5 = R_7 = 100$ kilohms; $R_3 = R_8 = 22$ kilohms; $E_1 = 16$ v; $E_2 = 22$ v.

Transistorised power amplifiers can use either single-ended or push-pull arrangement. Fig. 326 shows the circuit diagram of a push-pull earthed-base amplifier, which resembles the electron valve circuit designed for the same purpose. In the transistorised circuit, resistor R serves to obtain the positive bias voltage for the emitters. Push-pull circuits can also employ earthed-emitter stages. This circuit is sometimes preceded by a phase-inverting stage, which takes the place of input transformer T_1. Negative feedback may be used to decrease distortion.

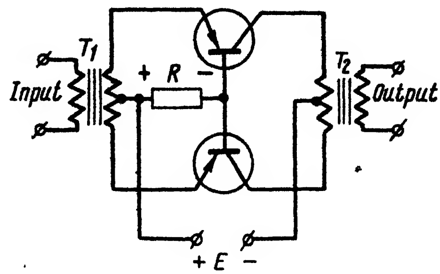


Fig. 326. The circuit diagram of a push-pull transformer-coupled amplifier

Fig. 327 gives the circuit diagram of a two-stage amplifier designed to reproduce speech and music from gramophone records. The amplifier is small enough to be fitted in a portable record-player. The first earthed-emitter stage is a voltage amplifier. This stage drives the push-pull output stage which is also of the earthed-emitter variety. The specifications of this circuit are: $R_1=10$

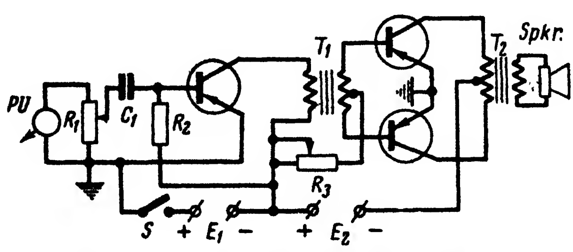


Fig. 327. The circuit diagram of a two-stage amplifier designed to operate from a record player

kilohms; $R_2=0.1$ megohms; $R_3=25$ kilohms; $C_1= 1$ mfd; $E_1=6$ v; $E_2=12$ v. R_3 is adjusted to provide the best amplification.

As we already know from the previous discussion, the properties of transistors vary with temperature changes. An increase of temperature decreases the resistance of the collector boundary, i.e., leads

to an increase of collector current, decreases the current gain coefficient and also decreases the input impedance. The collector characteristics are thus changed and the correct operating condition of the transistor is upset. These undesirable effects are counteracted by various methods of temperature stabilisation (or temperature compensation). In these methods, certain stabilising components are connected into individual stages of a transistorised circuit, the said components acting to maintain a relative constancy of operating condition.

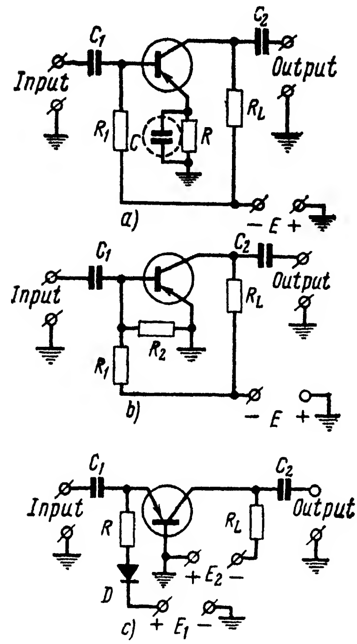


Fig. 328. Temperature stabilisation circuits employed by transistorised amplifiers

The stabilising method employed in the earthed-emitter circuit (which is most sensitive to temperature changes) consists in the application of a d.c. negative feedback circuit. This negative feedback circuit is provided by inserting resistor R, shunted by high capacitance, into the emitter lead (Fig. 328*a*). The value of R is made about 20-30% of the R_L value. As the temperature increases, both the collector current and the current flowing through R also increase. Then the voltage drop across R will become greater. Since this voltage drop acts as an additional negative bias for the emitter (with reference to the collector), the collector current will be reduced. Another way of improving the temperature stability of an earthed-emitter circuit is shown in Fig. 328*b*. Here, voltage divider R_1R_2 is employed to feed positive bias voltage to the base (i.e., negative voltage to the emitter).

Still another method of temperature stabilisation is shown in Fig. 328*c*. In this case, we deal with an earthed-base amplifier, where the reverse resistance of a semiconductor diode, type ДГ-Ц1, is connected into the emitter supply lead. As the temperature increases, the value of this reverse resistance is reduced, which increases the emitter positive bias. This effect stabilises the magnitude of amplification.

Although desirable, the temperature compensation is not absolutely necessary in many types of apparatus. This will be understood from the following. In many radio devices — for instance, in radio receivers — there is always a certain reserve of amplification. Should the thermal stability of such a receiver change sufficiently to decrease

the level of an incoming signal, a slight adjustment of the manual gain control will restore the required output volume. Hence, in such cases it is hardly worthwhile to complicate the receiver circuitry by the inclusion of special temperature-compensating components.

High-Frequency Amplifiers

The circuit of a transistorised high-frequency amplifier stage is shown in Fig. 329. This is an earthed-base version, with transformer output, in which the tuned circuit is connected directly into the collector lead. The bias voltage is fed to the emitter automatically from resistor R_1, the latter being shunted by capacitor C_1. The R_2C_2 link is a decoupling filter. Apart from its filtering functions, capacitor C_2 also plays the role of a blocking capacitor, allowing us to earth — with reference to a.c. — the rotor of the variable capacitor directly to the chassis.

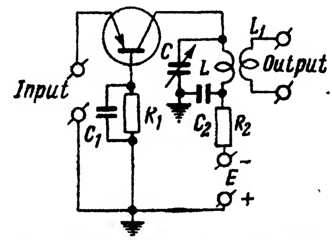


Fig. 329. The circuit diagram of a high-frequency amplifier

An interesting version of a high-frequency amplifier circuit is shown in Fig. 330. In this circuit arrangement, the input resistance of the transistor is connected in series with the tuned input circuit L_1C_1. This is possible only because of the low value of the input resistance of the transistor and cannot be done in a valve circuit. The oscillatory circuit L_2C_2 is connected into the collector circuit as an

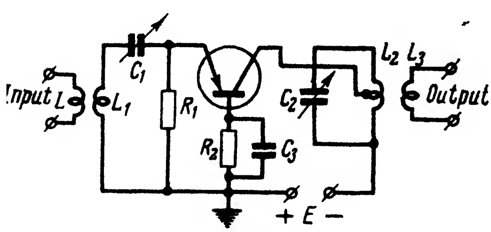


Fig. 330. The circuit diagram of a high-frequency amplifier with series-tuned circuit in the input

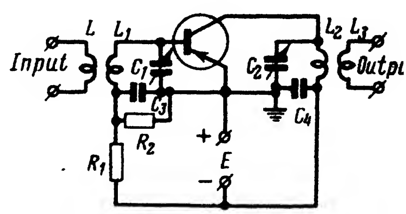


Fig. 331. The circuit of a high-frequency amplifier with the earthed-emitter arrangement

autotransformer. Such a method of connection assures impedance matching and is employed quite frequently. The automatic biasing voltage for the emitter is developed in the R_2C_3 link and is fed through resistance R_1.

The circuit of another high-frequency amplifier, in which the earthed-emitter arrangement is used, is shown in Fig. 331. In this

circuit, the emitter voltage is developed by voltage divider R_1R_2. Link R_1C_3 is a decoupling filter.

High-frequency amplifier circuits described above are equally applicable in intermediate-frequency amplifiers. Transistorised i.f. amplifier stages are usually coupled to each other through band filters, otherwise known as i.f. transformers.

Detectors and Mixers

Because of the non-linearity of transistor characteristics, a transistor circuit may be easily arranged to detect, as well as to amplify a.c. signals. Transistorised detector circuits may be made to operate in a similar manner to valve detector circuits intended for anode

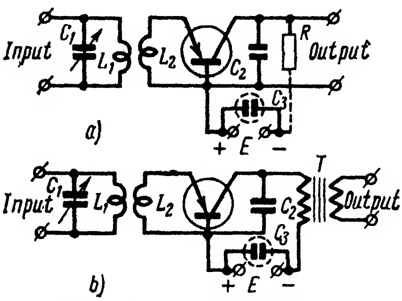


Fig. 332. The circuit diagrams of detector stages with the earthed-base arrangement

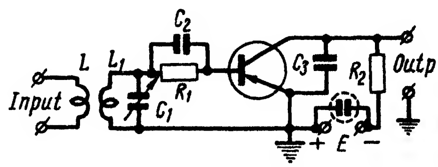


Fig. 333. The circuit diagram of an earthed-emitter detector stage

detection. Fig. 332 shows two such transistorised detector circuits with earthed bases. In the first version (Fig. 332*a*), the transistor works into a pure ohmic load, while in the second version (Fig. 332*b*) the detector load is represented by the primary winding of a low-frequency transformer. In both of these versions, capacitor C_2 bypasses the high-frequency component of collector current. Capacitor C_3 shunts the power supply source and must be of large capacitance. Similar circuits can also employ the earthed-emitter arrangement. A transistorised detector circuit, analogous to the grid-detector valve circuit, can also be devised. Such a circuit, employing an earthed-emitter arrangement, is shown in Fig. 333.

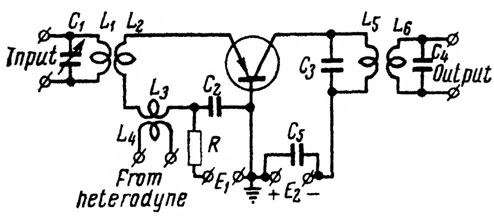


Fig. 334. The circuit diagram of an earthed-base mixer

When it is desired to convert frequency, any detector stage may be used as a mixer, if the input circuit of such a stage is supplied

h the incoming signal and with the signal developed by the hetero-dyne. Fig. 334 shows the circuit of such a mixer, in which both signals are applied to the emitter. In this arrangement, the mixer is inductively coupled to the heterodyne. The primary winding of the first i.f. transformer is connected into the collector circuit and is tuned to the intermediate frequency.

It is a better practice, however, to feed the incoming signal and the heterodyne signal to separate electrodes of a mixer, for instance—to feed one of the signals to the base and the other to the emitter. The circuit diagram of such a mixer is shown in Fig. 335. As may be seen, in this circuit the heterodyne signal is fed through capacitor C to choke Ch connected into the emitter lead. However, an alternative arrangement may be used here: coupling the mixer stage to the heterodyne inductively, as shown in Fig. 334.

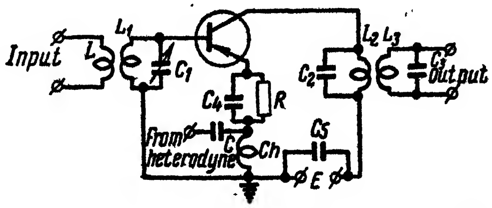


Fig. 335. The circuit diagram of a mixer with common emitter

Oscillators

Transistorised oscillators can employ a wide variety of circuit arrangements. Such oscillators may be used in radio transmitters. They have also found application in the heterodyne stages of radio receivers and in measuring equipment. When required, transistorised oscillators can be used to generate signals in the audible range.

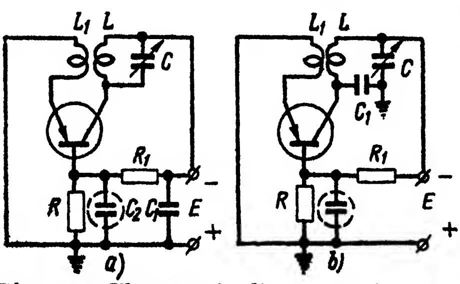


Fig. 336. The circuit diagrams of oscillators with inductive feedback

These oscillators may be designed to generate either nearly sinusoidal signals, or else such signals whose shape is sharply non-sinusoidal. (Sawtooth or rectangular signal oscillators are known as relaxation oscillators.) We shall discuss here only such oscillators which develop sinusoidal signals. These oscillators usually employ junction transistors and are frequently based on well-known valve-oscillator circuits.

Fig. 336*a* gives the circuit diagram of a transistorised oscillator employing inductive feedback and earthed base. The stage operates from a single power supply. The positive bias necessary for the emitter circuit is derived from resistor R connected into the base

lead. The voltage drop is built up across this resistor by virtue of the base current and of the current flowing through resistor R_1. The optimum operating condition of the stage is set by resistor R, which should be made variable for this purpose. If it is required to earth the rotor of tuning capacitor C, the described oscillator circuit should be rearranged as shown in Fig. 336*b*. The output voltage developed by the oscillator may be taken off directly between the collector and earth. Alternatively, an inductive arrangement may be introduced into the circuit for the purpose of picking up the output energy.

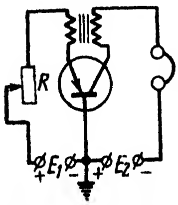


Fig. 337. The circuit diagram of the simplest transistorised audio-frequency oscillator

The simplest circuit of a transistorised audio-frequency oscillator, employing inductive feedback, is shown in Fig. 337. Here, the feedback is provided by an ordinary low-frequency transformer and, hence, the oscillatory circuits are formed by the inductance and distributed capacitance of the transformer windings. The oscillator is powered by two small batteries, one of the batteries providing the emitter voltage of about 1.5 v, and the other—the collector voltage, which may be anything from 3 to 15 volts. In order to make the stage oscillate, the 15,000-ohm rheostat R should be adjusted. This rheostat is provided for the purpose of obtaining oscillations at the minimum emitter current. Should it happen that the stage fails to oscillate at any position of the pointer of rheostat R, the reversal of the two leads of one of the transformer windings will usually cure the trouble and the stage will break into oscillation. A pair of earphones shown connected into the collector lead of the stage is used for monitoring and will reproduce an audible sound when the oscillations begin. If desired, the earphones may be replaced by a small loudspeaker connected through an output transformer.

Two versions of transistorised oscillator circuits are shown in Fig. 338, both of these versions using inductive feedback and earthed-emitter arrangement. Although these circuits are shown provided with tuned circuits and, hence, are intended for high-frequency service, they will operate equally well in the low-frequency range and can be used as alternative types of simple audio oscillators. For the latter type of duty it is only necessary to replace tuned circuit LC and coil L_1 by a low-frequency transformer, leaving unaltered the remaining circuit components shown in Fig. 338.

Transistorised oscillators employing three-point autotransformer-type and capacitive feedback circuit arrangements are shown in Fig. 339. These oscillators use earthed-emitter arrangements and require only one power supply source for their operation. Fixed resistors in the circuits of these oscillators determine the correct operating conditions of the stages and must be, therefore, properly

selected. Such selection is also necessary because of another consideration; the values of these resistances also determine the shape of oscillations. In the circuit given in Fig. 339*b*, capacitance C_2 must be about 10 times as great as capacitance C_1.

Earthed-base oscillators employing point-contact transistors should be regarded as a special group of transistorised oscillators.

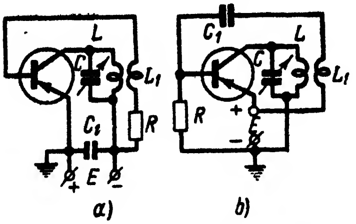


Fig. 338. The circuit diagrams of oscillators with inductive feedback and earthed emitters

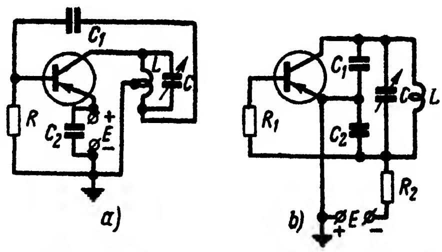


Fig. 339. The circuit diagrams of transistorised oscillators employing autotransformer-type feedback (*a*) and capacitive feedback (*b*)

Oscillations are obtained in these circuits by connecting into the base lead a high-value additional feedback resistor in series with the natural base resistance r_b of the transistor. Because of the additional resistor, the stability of the stage is upset and the circuit breaks into oscillation. In radio literature, these oscillators are frequently called negative-resistance oscillators.

Fig. 340 gives several circuit versions of such oscillators. In the circuit shown in Fig. 340*a*, a series-resonant tuned circuit is connected into the emitter circuit. Here, ohmic resistor R provides negative

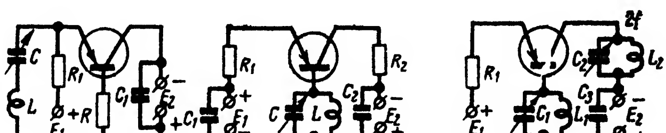


Fig. 340. Several versions of negative-resistance oscillator circuits

feedback. Since the circuit employs a point-contact transistor, $i_c > i_e$. The flow of current i_c through resistances $r_b + R$ builds up a voltage drop which coincides in phase with the voltage across tuned circuit LC, and, thus, a positive feedback is obtained. The value of feedback and the operating condition of the oscillator are adjusted by varying resistors R_1 and R.

The oscillator circuit shown in Fig. 340*b* employs a parallel-resonant tuned circuit as the feedback impedance. Like the feedback

resistor in Fig. 340*a*, this tuned circuit is also connected into the base lead. Various shapes of oscillations may be secured in this oscillator by appropriate adjustment of resistors R_1 and R_2. Non-sinusoidal oscillations are sometimes generated for frequency-multiplication purposes. A frequency-doubling circuit is shown in Fig. 340*c*. In this circuit, the oscillations of fundamental frequency f are generated in tuned circuit L_1C_1, while tuned circuit L_2C_2, adjusted to frequency $2f$, separates the second harmonic.

If a quartz crystal is used in place of the fundamental tuned circuit, a highy-stable crystal-controlled transistorised oscillator circuit will result. Such precision oscillators are commercially available and are noted not only for their high performance, but also for the extreme economy in power consumption; an oscillator of this type will operate from a single 1.3-v torchlight cell and will consume only 100 microamperes of current. (Compare this to the power consumed by the miniature bulb of the usual pocket-type torchlight; the bulb is usually designed to operate from about 4.5 volts and draws approximately 300 milliamperes, i.e., 300,000 microamperes!—*Translator's note.)*

It should be noted that point-contact transistors, when used in oscillators, can generate signals whose frequency may be as high as several hundred megacycles. The upper frequency range of junction transistors does not exceed some dozen megacycles.

Radio Receivers

The simplest circuit of a transistorised two-stage straight-amplification radio receiver (type 0-V-1) is shown in Fig. 341. The receiver circuit comprises two junction transistors, employing earthed-emitter arrangement. The circuit specifications are as follows: $R_1 = 25$ kilohms; $R_2 = 0.3$ megohms; $C_1 = 8$ mfd; $E = 30$ v.

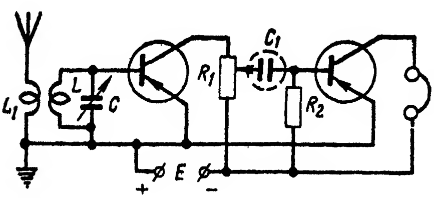


Fig. 341. The circuit diagram of the simplest transistorised radio receiver type O-V-1

The circuit of a three-stage straight-amplification receiver (type 1-V-1) is shown in Fig. 342. This receiver employs the earthed-base connection of transistors. The circuit diagrams of individual stages of the given receiver were given in Fig. 326, Fig. 330 and Fig. 332*b*. Because of this, no further discussion of these stages is required. The *RC* unit is a decoupling filter.

Regenerative receivers, based on transistors, can also be designed. An example of such receiver is given in Fig. 323. This is a three-stage receiver, known as type 0-V-2. The first stage employs a

point-contact transistor, the regenerative tuned circuit being connected into the base lead. The stage can be made regenerative by an appropriate adjustment of variable feedback resistor *R*. Low-frequency amplifier stages use transformer coupling and the earthed-base arrangement.

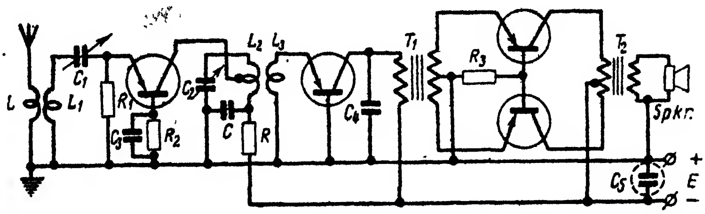


Fig. 342. The circuit diagram of radio receiver type 1-V-1

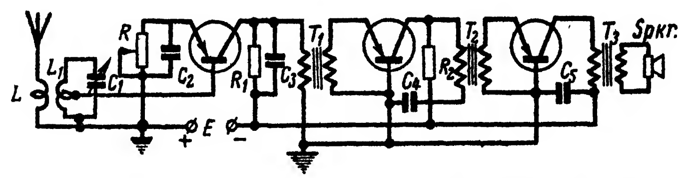


Fig. 343. The circuit diagram of type O-V-2 regenerative radio receiver

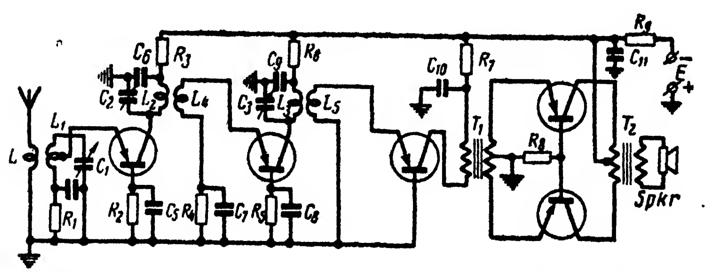


Fig. 344. The circuit diagram of type 2-V-1 radio receiver

The circuit diagram of a four-stage straight-amplification receiver (type 2-V-1) is shown in Fig. 344. All the stages in this set use the earthed-base arrangement. Since the input impedance of such stages is low, the first tuned circuit is autotransformer-coupled to the emitter circuit of the first transistor. The emitters are automatically biased in the high-frequency stages and also in the push-pull output stage. Decoupling filters are employed in collector circuits of all the stages in order to suppress various types of parasitic coupling. A similar filter is also provided in the common supply lead.

One of the many possible circuit arrangements of transistorised superheterodyne receivers is shown in Fig. 345. With the exception of the heterodyne stage, the receiver stages employ the earthed-emitter arrangement. The circuit is quite simple and uses junction transistors. Although a loudspeaker is shown connected to the output stage of the receiver, one additional i.f. stage and one additional l.f.

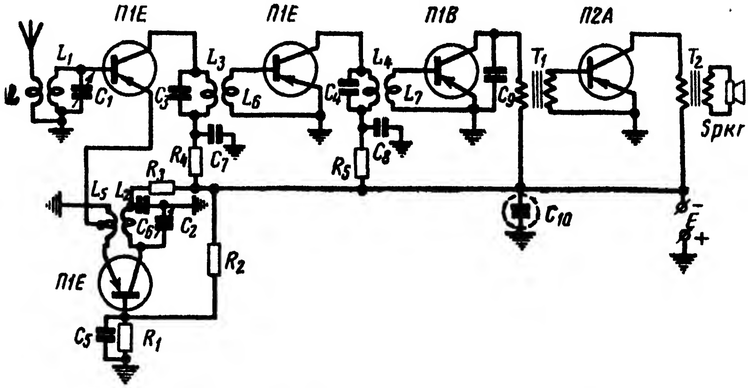


Fig. 345. The circuit diagram of a simple superheterodyne radio receiver

stage should be added to the circuit, if a really good loudspeaker reception is desired. Decoupling filters are connected into the collector circuits of the mixer, heterodyne and i.f. amplifier. The heterodyne stage employs inductive feedback and is autotransformer-coupled to the mixer. The positive bias to the heterodyne emitter is fed from a voltage divider consisting of resistors R_1 and R_2. The i.f. transformers used by the set should be of the step-down variety.

139. HOW TO HANDLE SEMICONDUCTOR DEVICES

Semiconductor diodes and triodes can be connected to their associated wiring by means of soldering, plug-and-socket connectors or screw terminals. When soldering is resorted to, special care should be taken not to damage the highly temperature-sensitive devices by the heat developed by the soldering iron in the process of soldering. A few words of precaution are in order in this connection.

When soldering the terminal leads of a semiconductor device to the associated external wiring, never touch the leads with the soldering iron at points located nearer than 10 mm to the body of the semiconductor device. The soldering iron should be of the low-power variety (50-60 watts), and, when applied to a lead, should not be held down to it for longer than 2-3 seconds, after which it

should quickly be moved away from the lead to give the lead a chance to cool off. The melting temperature of the solder should not be higher than 150°C. It is recommended to insert a piece of metal between the body of the semiconductor device and the point on the lead to which the soldering iron is applied; the metal will help to dissipate the heat generated by the soldering iron and will lessen the possibility of propagation of this heat to the semiconductor. The body of the soldering iron should be thoroughly insulated from its heating element.

Besides the above precautions, the semiconductor must be also protected from mechanical damage in the process of installation. Although semiconductor devices are quite robust, their terminal leads can be easily damaged, particularly by bending. Because of this, when bending a lead, do it not closer than 5 mm to the semiconductor body. When dealing with transistors type C1 or type C2 remember that the terminal leads of these devices must not be bent, or soldered. Plug-and-socket connectors or screw terminals are the only permissible means of attaching the terminals of C1 and C2 transistors to their circuits. When dealing with any type of semiconductor devices, do not suspend such devices from their associated wiring, as this would lessen the resistance of the devices to mechanical shocks. Never install semiconductors close to hot components and, wherever possible, provide them with metal ribs or with other types of heat-dissipating contrivances.

When operating a semiconductor device in an electric circuit, never exceed the maximum permissible current and voltage ratings assigned to the device. Point-contact devices should not be subjected to electrical overloads even for very short periods of time. If this rule is not observed, the devices will be either damaged or else their parameters will be permanently altered to such an extent as to make the semiconductors unserviceable. When applying power to a transistor, make certain that the base is energised first. When handling point-contact transistors, make sure that the ohmic resistance of the emitter circuit has a value of at least 500 ohms. If this is not done, the transistor operation may become unstable and the device will be subjected to electrical overloads.

140. QUESTIONS AND PROBLEMS

1. What are the advantages and disadvantages of semiconductor devices, as compared to electron valves?
2. What is the electron-type conductivity of semiconductors?
3. What are the holes in a semiconductor and how do they differ from positive ions?
4. When electric current flows through a semiconductor with hole-type conductivity, what particles are transferred during the process?
5. How does the rectification of current take place at the boundary of two semiconductors possessing different types of conductivity?

6. What causes the reverse current flow in a semiconductor rectifier?

7. How do the properties of point-contact and junction semiconductor diodes differ?

8. A semiconductor diode has the following parameters: $R_f = 50$ ohms; $R_r = 100$ kilohms; $C = 40$ pf. At what frequencies will the rectifying action of such a diode be noticeably impaired by the shunting effect of its own distributed capacitance?

9. Why is it inconvenient to show to the same scale the curves for forward and reverse currents on the volt-ampere characteristic of a semiconductor diode?

10. Draw various circuits of rectifiers employing semiconductor diodes.

11. Why will a transistor refuse to function if the distance between the emitter and collector boundaries is made excessive?

12. What is the difference and what is the similarity between the operation of a valve triode and a transistor?

13. Draw the circuit diagrams of amplifier stages with earthed-base, earthed-emitter and earthed-collector arrangements, when the transistor employing these stages is of the *n-p-n* type. Show all the paths of various currents in these circuits.

14. Given a transistorised amplifier stage with the earthed-base arrangement. In this stage, the collector circuit is supplied with power from a source with voltage $E_2 = 40$ v. The resistance values for different parts of this circuit (in respect to direct current) are, respectively: $R_b = 50$ ohms; $R_c = 10$ kilohms; $R_L = 10$ kilohms. When the emitter voltage is increased by 0.2 volts, resistance R_c is decreased by 4,000 ohms. Neglecting the value of R_b, determine the voltage gain coefficient of the stage.

15. When the emitter current is changed by 2 ma in a certain junction transistor, the collector current varies by 1.85 ma. What is the value of the current gain coefficient in such a stage?

16. What is meant by the undesirable capacitive influence of the collector boundary?

17. What are the advantages and disadvantages of point-contact transistors as compared with junction transistors?

18. Draw a family of emitter characteristics, and a family of collector characteristics, of a transistor and explain their meaning.

19. What are the basic properties of a transistorised amplifier with earthed base?

20. What are the advantages of an amplifier with earthed emitter?

21. Why is it that an amplifier with earthed collector may be called an emitter follower?

22. Devise the circuit diagram of a transistorised two-stage low-frequency amplifier.

23. Devise the circuit diagram of a transistorised type 1-V-1 radio receiver.

24. How does the temperature affect the properties of semiconductor diodes and triodes?

25. What methods of temperature stabilisation may be used in a transistorised amplifier stage?